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Thyristor Speed Control Circuit Design for D.C. Shunt Motors Supplied from A.C. Mains

J. MERRETT
Mullard Central Application Laboratory

Three methods of electronically controlling the speed of d.c. shunt motors are described in this article. Full design details and practical circuits are given for two control systems, each with 1% regulation, using either a tachometer or an armature back e.m.f. bridge. A simpler circuit is also described with a regulation of between 5% and 10%. A speed range of over 100 : 1 is possible with the more complex circuits and of over 10 : 1 with the simpler circuit; all have good regulation at low speed. Current limit and integral starting circuits are included in all systems and p.t.c. overload protection may also be incorporated, if required.

INTRODUCTION

When a variable speed electric motor has been required, it has been customary in the past to use a d.c. shunt wound machine and either vary the field current or the armature voltage. Such systems used either field regulator resistors or Ward Leonard sets for armature control. Thyratrons and saturable reactors have also been used for armature control. In most cases the equipment required was bulky and expensive so that these electrical methods of speed control were only used where the alternative mechanical methods were unsuitable. Now that relatively cheap, reliable thyristors are available it is possible to control the speed of d.c. shunt motors, using the variable armature voltage technique, directly from the single or three phase a.c. supply at a cost which is competitive with alternative systems. This article, together with earlier articles (Refs. 1, 2), provides an introduction to some of these techniques.

Design procedures for motor speed control are described and illustrated by some worked examples showing the methods adopted when designing practical circuits. The performance of these circuits is shown to be as good as, or better than, can be achieved by other methods. In addition, it is possible to provide many auxiliary features, such as current limit circuits, starting circuits, open circuit field protection, field stabilisation, or p.t.c. thermistor overload protection. Some of these could not easily be incorporated in the earlier control system.

ELEMENTARY SYSTEM DESIGN

The block diagram of a typical electronic speed control system is shown in Fig. 1, from which the principle used will be evident. A signal proportional to motor speed is compared with a reference signal; the difference or error signal is amplified and used to control the thyristor trigger angle in such a way as to reduce the speed error, and so maintain speed nearly constant.

If sufficient motor data is available, and the supply voltage is known, it is possible to estimate the thyristor firing angles for a range of motor speeds and load torques. Similarly, data for the thyristor firing circuit will enable the designer to estimate the magnitude of the control voltage which is required to obtain a particular firing angle. We can now derive a direct relationship between motor speed and load torque and the input control voltage to the firing circuit. If motor speed is to be held between certain prescribed limits for a certain change in load, the control signal required to do this can then be calculated. Once the sensitivity of the speed measurement system has been established, the required amplifier gain can be estimated. The circuit performance at other operating points can then be checked by a similar procedure and modified, if necessary, to bring it within the required specification.

To illustrate the above procedure more clearly, an example will be followed through in the order outlined above. In the course of this example reference will be made to design graphs which are fully described in Refs. 1, and 2 but for convenience these are reproduced in this article as Figs. 2a, 2b, 2c and 8.

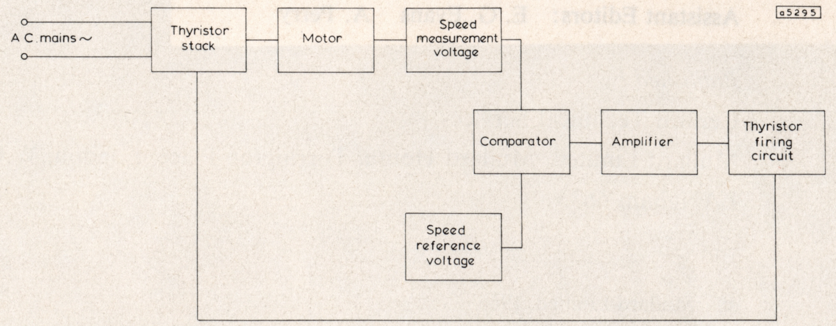

Fig. 1—Block diagram of motor speed control system

For the purpose of this example, assume that the motor being used is a 1500 rev/min 5hp machine with the following electrical specification.

Armature voltage = 180V d.c.
Armature full-load current = 25A d.c.
Armature resistance = 0·36Ω
Armature power factor (cos ϕ) = 0·4

The supply voltage is 240V r.m.s.

The motor operating parameters are therefore as follows:

$$\text{Armature stalled current} = \frac{180}{0{\cdot}36},$$
$$= 500\text{A}.$$
$$\text{Torque factor at full-load} = \frac{25}{500},$$
$$\text{that is, } T_f = 0{\cdot}05.$$
$$\text{Speed factor (a) at full-speed} = \frac{180}{240\times\sqrt{2}},$$
$$= 0{\cdot}53.$$

From Fig. 2b it will be seen that at full speed and full load, that is, when $a = 0{\cdot}53$ and $T_f = 0{\cdot}05$, the thyristor firing angle will be approximately 95°. At the same speed but reduced load, say at a torque factor of 0·01, the thyristor firing angle will be about 120°. If the speed is to be kept constant while the load changes from 1/5 load to full load the thyristor firing angle must therefore be changed from 120° to 95°.

Before proceeding further it will be useful to derive the design equation relating speed regulation with control loop gain. Let,

firing angle at minimum load = xf_1 degrees
firing angle at maximum load = xf_2 degrees
trigger circuit input sensitivity = θ degrees/volt

Then, if speed is to be maintained constant, change in input voltage (δV_F) to the trigger circuit is given by:

$$\delta V_F = \frac{(xf_1 - xf_2)}{\theta}$$

Now, if the tachometer sensitivity = V volts per 1000 rev/min

and motor speed = N rev/min

then tachometer output voltage (V_T) is given by the equation,

$$V_T = \frac{V.N}{1000}\text{ volts.}$$

Change in tachometer voltage when speed changes by δN is given by the equation,

$$\delta V_T = \frac{V}{1000}\delta N.$$

But when the feedback loop is closed

$$\delta V_F = G \times V_T,$$

where G is the error-amplifier gain.

Unless G is very large the motor speed must change by a measurable amount to satisfy this condition. If the motor speed is allowed to fall, then,

$$\delta V_F < \frac{(xf_1 - xf_2)}{\theta}.$$

The values of xf_1 and xf_2 are however not known to great accuracy as they are obtained from the design curves with, at best, an accuracy of ±5%. The value of δV_F will be reduced by approximately the same fraction as the fractional fall in motor speed. If this is small we can then write, with little error,

$$\delta V_F \simeq \frac{(xf_1 - xf_2)}{\theta}.$$

We then obtain,

$$\frac{(xf_1 - xf_2)}{\theta} = G \times \frac{V.\delta N}{1000},$$

or

$$G = \frac{(xf_1 - xf_2)}{\theta} \times \frac{1000}{V.\delta N}.$$

$$\text{Motor regulation} = \frac{\delta N}{N} \times 100\%.$$

$$\text{Therefore, regulation} = \frac{(xf_1 - xf_2)}{\theta} \times \frac{1000}{V.N.G}.$$

In the practical example referred to above, assume that these constants have the following values:

$xf_1 = 120°$
$xf_2 = 95°$
$\theta = 18°$/volt
V = 10V per 1000 rev/min
N = 1500 rev/min
G = 10

Inserting these values into the equation for regulation gives,

$$\text{Regulation} = \frac{(120-95)}{18} \cdot \frac{1000}{10.1500.10} \cdot 100\%$$

$$= 0{\cdot}93\%.$$

It will be seen from this example that at high speed good motor regulation is achieved with only a low loop gain. At lower speeds however higher gain will be necessary if good regulation is required, as shown by the following example.

Taking the same example but with the motor operating at 1/10 full-speed, that is, $a = 0{\cdot}053$, then,

$xf_1 = 162°$
$xf_2 = 140°$

$$\text{Regulation} = \frac{22}{18} \cdot \frac{1000}{10.150.10} \cdot 100\%$$

$$= 8{\cdot}1\%.$$

By rearranging the equations, it is clearly possible to calculate the minimum loop gain required to achieve a given regulation.

It is possible to design a number of different speed control circuits based upon the general principles which have been described; they will differ in their methods of speed measurement and in their general complexity. Their complexity will depend upon the degree of control and other facilities required. By way of example, three different control systems are described and analysed below.

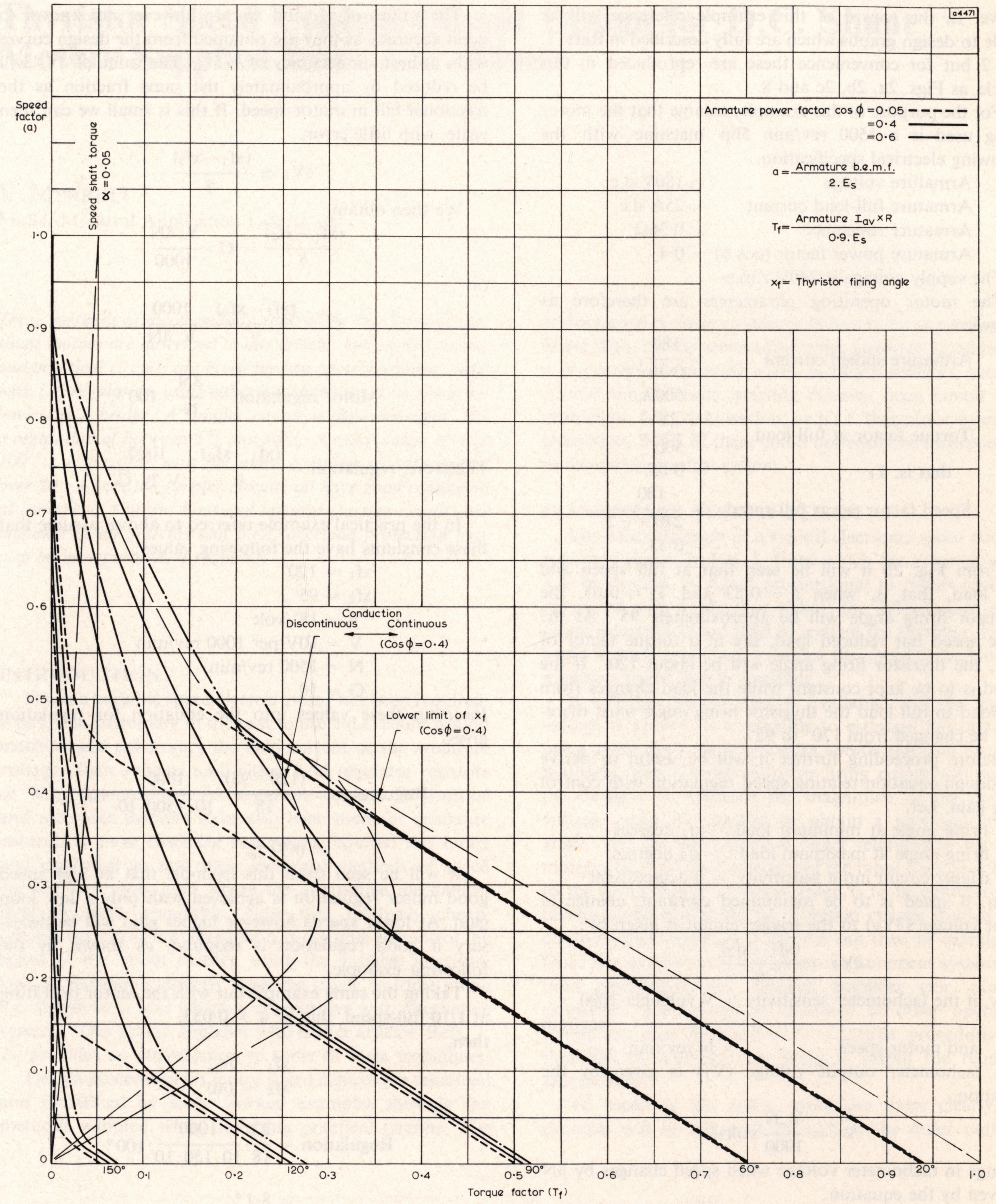

Fig. 2a—Curves for derivation of firing angle, using speed factor, torque factor and cos ϕ

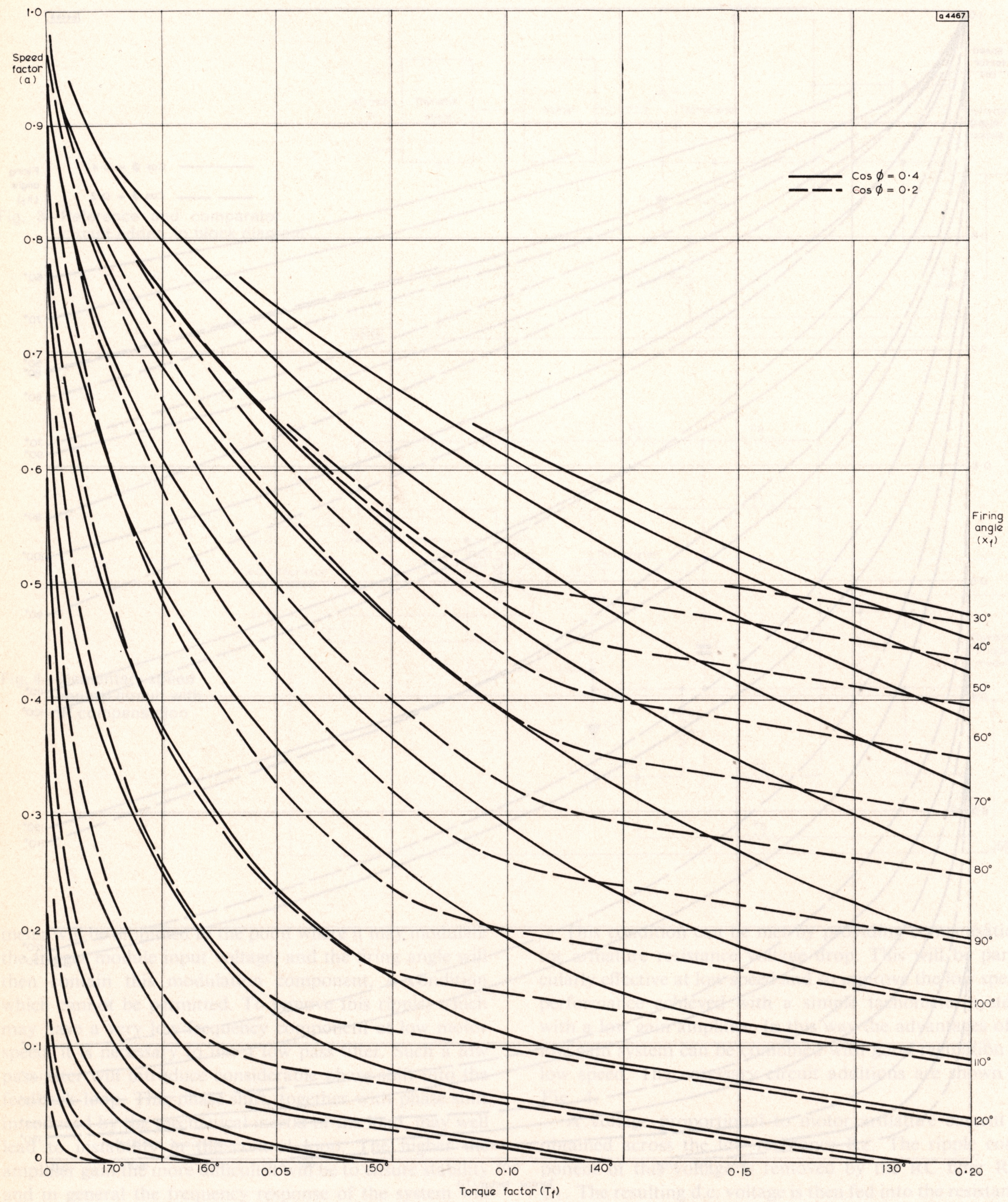

Fig. 2b—Curves for derivation of firing angle, using speed factor, torque factor and cos ϕ

TACHOMETER SPEED CONTROL

The tachometer speed control system is one of the most complex circuits in terms of numbers of components; the various functions are, however, performed by separate parts, and it is therefore the easier circuit to design and analyse since each part can be considered in isolation.

The simplest possible circuit is that shown in block form in Fig. 1. It will be convenient to build up a practical circuit from this block diagram, stage by stage. The first of these might take the form shown in Fig. 3. Here a variable reference voltage is provided by the voltage regulator diode (zener) D_1 and potentiometer R_4. This voltage is added in reverse to the output voltage from the d.c. tachometer in a resistive adding network R_1, R_2, R_3, so that the input to the amplifier is the algebraic difference between these two voltages. The motor speed will adjust itself so that the output voltage of the tachometer is very near to the reference voltage provided by R_4. The difference voltage, when amplified by the d.c. amplifier, must be just sufficient to provide the voltage input to the trigger circuit to cause it to give the required thyristor trigger angle to satisfy the motor operating conditions. The

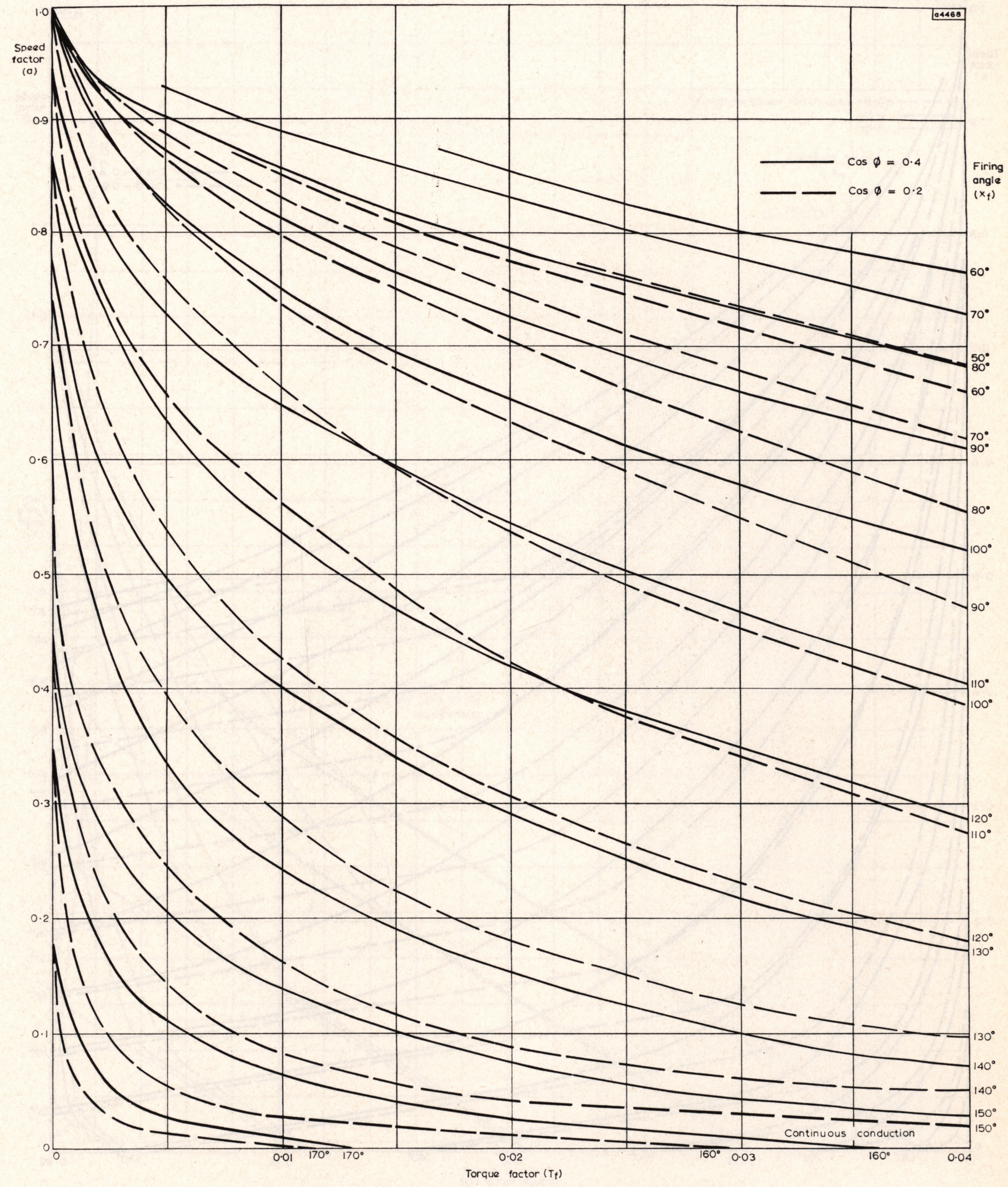

Fig. 2c—Curves for derivation of firing angle, using speed factor, torque factor and cos ϕ

required trigger angle can be read from Figs. 2a, 2b, 2c as explained earlier. A change in motor speed brought about by any external influence (for example, a change in mechanical load) will result in a change in difference voltage at the amplifier input which in turn will produce a change in trigger angle in the correct sense to maintain the speed nearly constant.

It has already been shown that to achieve good regulation at the low speed end of a wide speed range system of the above type it is necessary to have a higher amplifier gain than is necessary at maximum speed. If the amplifier gain is high, any residual ripple voltage from the tachometer will be amplified to the point where it may modulate the trigger module input voltage, and the firing angle will then contain this modulation component, a condition which cannot be permitted. To remove this ripple, which may have a very low frequency component at low motor speed, it is necessary to use a low pass filter. Such a low pass filter will introduce considerable phase shift into the feedback loop. This phase shift, together with phase shift introduced by the mechanical inertia of the load, may well lead to instability in the control loop. The higher the amplifier gain the more difficult it will be to ensure stability and in general the frequency response of the system will have to be made very low for stable operation.

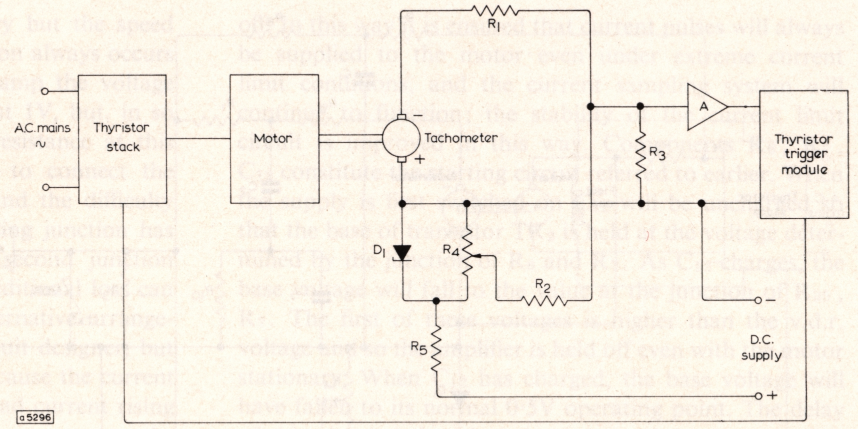

Fig. 3—Reference and comparator circuit added to block diagram

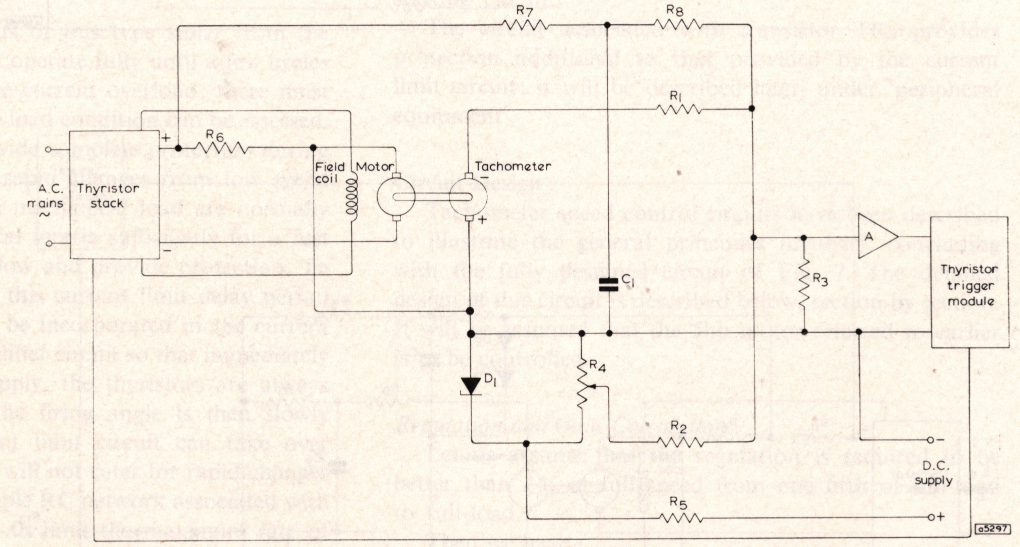

Fig. 4—Elementary speed control circuit with IR compensation

In many applications high absolute speed accuracy is not required over a very wide speed range. High accuracy may be required for example over a 3:1 speed range, but good regulation with load change may still be required at much lower speeds for machine starting or for initial setting up, but with a much relaxed accuracy requirement. High accuracy is used here to mean good long-term speed stability, while regulation is only related to speed change under load variations.

This condition can be met by providing compensation for armature resistance voltage drop. This will be particularly effective at low speed and so improve the low speed performance achieved with a simple tachometer system with a low gain amplifier. In this way the advantages of a low gain system can be combined with good regulation at low speed. The necessary circuit additions are shown in Fig. 4.

A voltage proportional to motor armature current is obtained across the low resistance R_6. The ripple component of this voltage is removed by the RC filter R_7, C_1. The resulting d.c. voltage is then fed into the resistance adder through R_8.

It will usually be necessary to provide some smoothing at the output of the tachometer to remove residual commutation ripple. In most cases a simple filter will introduce sufficient phase shift to make the control loop unstable. An RC network R_9, C_2, R_{10}, C_3 of the type shown in Fig. 5 will normally suffice to stabilise the loop. In this case, the resistor R_1 (in Fig. 3) is split into two parts R_1 and R_9 as shown (Fig. 5). One other additional component is shown in Fig. 5. This is the voltage depend-

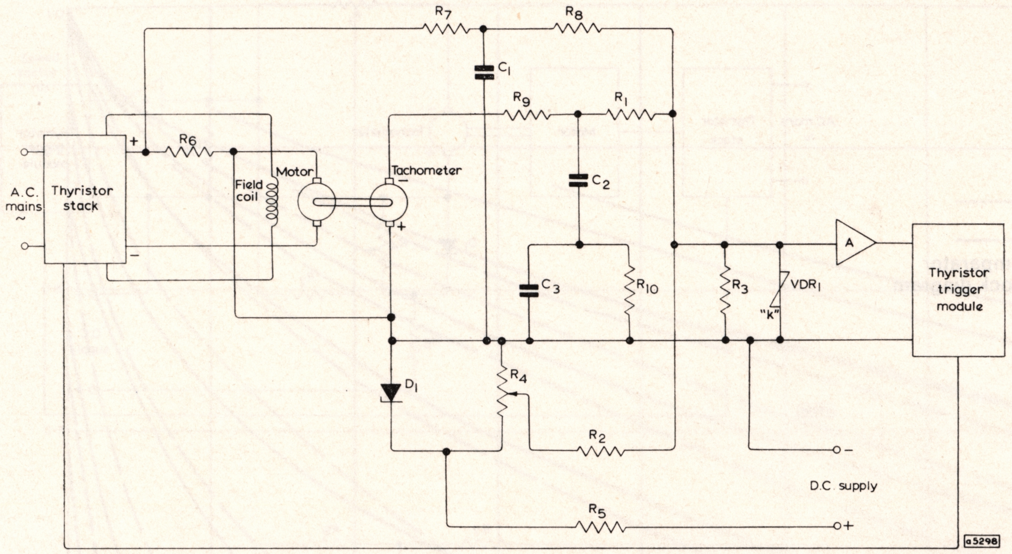

Fig. 5—Tachometer speed control circuit with filter and v.d.r. clamp

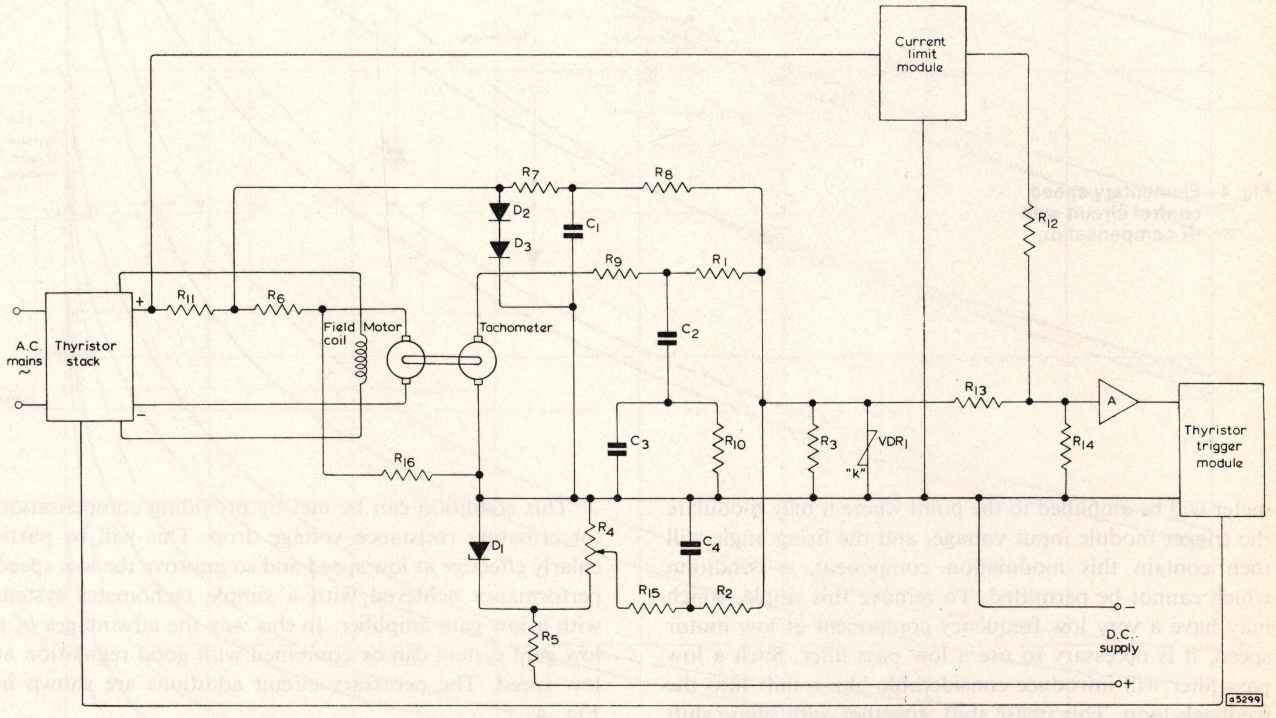

Fig. 6—Tachometer speed control with current limit

ent resistor VDR_1 in parallel with R_3. When the motor speed setting is changed by a large amount, by altering the setting of R_4, a large voltage will be applied to the amplifier input until the motor is able to change speed to equalise the voltages again.

VDR_1 serves to limit this voltage excursion to a value which will not damage the amplifier. It also serves another purpose which will be referred to later.

Most motor speed control systems will require some additional features; one of these will almost certainly be current limit. Broadly speaking, this consists of a current sensing element, either in the a.c. mains supply or in the motor armature circuit. A comparator is used to compare the current magnitude with a reference. When the current exceeds a predetermined value, a control voltage is fed to the trigger module in a sense which decreases the thyristor conduction period, thus maintaining the load current near its preset maximum value.

The additional circuit components required for current limit are shown in Fig. 6. Here R_{11} is used to measure the armature current and the comparator circuit is shown as a circuit module; the output from this module is connected into the adding junction. A certain difficulty arises because current limit action will be required under the conditions where the motor is stationary but the speed control is set to high speed. (This condition always occurs when starting from rest.) VDR_1 will clamp the voltage across R_3 to a low value, typically about 1V, but, in so doing, will impose a low incremental resistance at this point; it would, therefore, be difficult to connect the current limit output here. One way round the difficulty is shown in Fig. 6, where a second adding junction has been provided for current limit. This second junction increases the loop attenuation but the additional loss can easily be made up in the amplifier. Alternative arrangements will suggest themselves to the circuit designer, but the same problems are likely to arise, because the current limit function has not only to prevent load current rising with a decrease in load impedance, but also to counter the relatively large effect of the reduction of tachometer voltage at low speed, during either starting or heavy overload.

All current limit circuits of this type suffer from the limitation that they do not operate fully until a few cycles after the application of the current overload; there must be some current before the load condition can be assessed. They do not therefore provide complete protection during motor starting or during rapid changes from low speed to high speed. Changes in mechanical load are normally slowed down by mechanical inertia sufficiently for a fast current limit circuit to follow and provide protection. To provide protection during this current limit delay period a 'slow start' circuit may be incorporated in the current limit module or in the amplifier circuit so that immediately after switching on the supply, the thyristors are always fired late in the cycle. The firing angle is then slowly advanced until the current limit circuit can take over control. This arrangement will not cater for rapid changes in the speed control. A simple RC network associated with R_4 can, however, be used to limit the maximum rate of change of control voltage to a level at which the current limit circuit can follow and provide protection.

The components R_{15}, C_4 shown in Fig. 6 provide the required time constant. The resistors R_{15} and R_2 in Fig. 6 form the input series resistor of the adding junction, and replace R_2 in Fig. 5. A slow start circuit will be described later, together with the error amplifier. One further circuit detail which has been omitted in the discussion so far, but which improves circuit stability, is also shown in Fig. 6. It takes the form of a diode clamp (D_2, D_3, R_{16}) which is designed so that at full load current the clamp just becomes operative to prevent further IR compensation being applied in opposition to current limits.

The complete circuit diagram of a tachometer speed control system based on the above principles is shown in Fig. 7. In this circuit diagram the error amplifier is shown in detail. It will be seen to have a balanced input to reduce drift caused by temperature or supply voltage changes. The output is taken from a low gain stage with a low output impedance; there is overall negative feedback through R_{36} to stabilise the gain.

The output of the amplifier is clamped by the diodes D_{21}, D_{22} at the lower limit of its voltage excursion so that the trigger module can never be completely turned off. In this way it is ensured that current pulses will always be supplied to the motor even under extreme current limit conditions, and the current sampling system will continue to function; the stability of the current limit circuit is improved in this way. Components R_6, R_7, C_{12} constitute the starting circuit referred to earlier. When the supply is first switched on C_{12} will be uncharged so that the base of transistor TR_2 is held at the voltage determined by the junction of R_6 and R_7. As C_{12} charges, the base voltage will fall to the value at the junction of R_{35}, R_7. The first of these voltages is higher than the v.d.r. voltage and so the amplifier is held off even with the motor stationary. When C_{12} has charged, the base voltage will have fallen to its normal 0·5V operating point. The delay of this circuit is largely independent of the speed setting of R_{23}. This contrasts with the delay circuit R_{24} and C_5, which for this reason cannot be satisfactorily used as a starting circuit.

The circuit associated with transistor TR_3 provides protection additional to that provided by the current limit circuit; it will be described later, under 'peripheral equipment'.

Circuit Design

Tachometer speed control circuits have been described to illustrate the general principles involved, concluding with the fully designed circuit of Fig. 7. The detailed design of this circuit is described below, section by section. It will be assumed that the 5hp motor referred to earlier is to be controlled.

Regulation and Gain Calculation

Let us assume that the regulation is required to be better than 1% at full speed from one fifth of full-load to full-load.

Then we have,

$$\text{Regulation} = \frac{(xf_1 - xf_2)}{\theta} \cdot \frac{1000}{\text{V.N.G.}} \cdot 100\%.$$

Therefore, the value of gain (G) is given by the equation.

$$G = \frac{(xf_1 - xf_2)}{\theta} \cdot \frac{1000}{1.\text{V.N.}} \cdot 100.$$

Taking $\theta = 18°/\text{V}$, $N = 1500$ rev/min, $V = 10\text{V}$ per 1000 rev/min,
then

$$G = \frac{(xf_1 - xf_2)}{18} \cdot \frac{1000}{10.1500} \cdot 100.$$

Therefore

$$G = 0{\cdot}37(xf_1 - xf_2).$$

From Fig. 2b, we may obtain that $xf_1 = 120°$ and $xf_2 = 95°$.
Therefore

$$G = 0{\cdot}37 \times 25$$

$$= 9{\cdot}25.$$

A practical value to which to design would therefore be a gain of 10, measured from the adder input to the amplifier output at its point of connection to the trigger module.

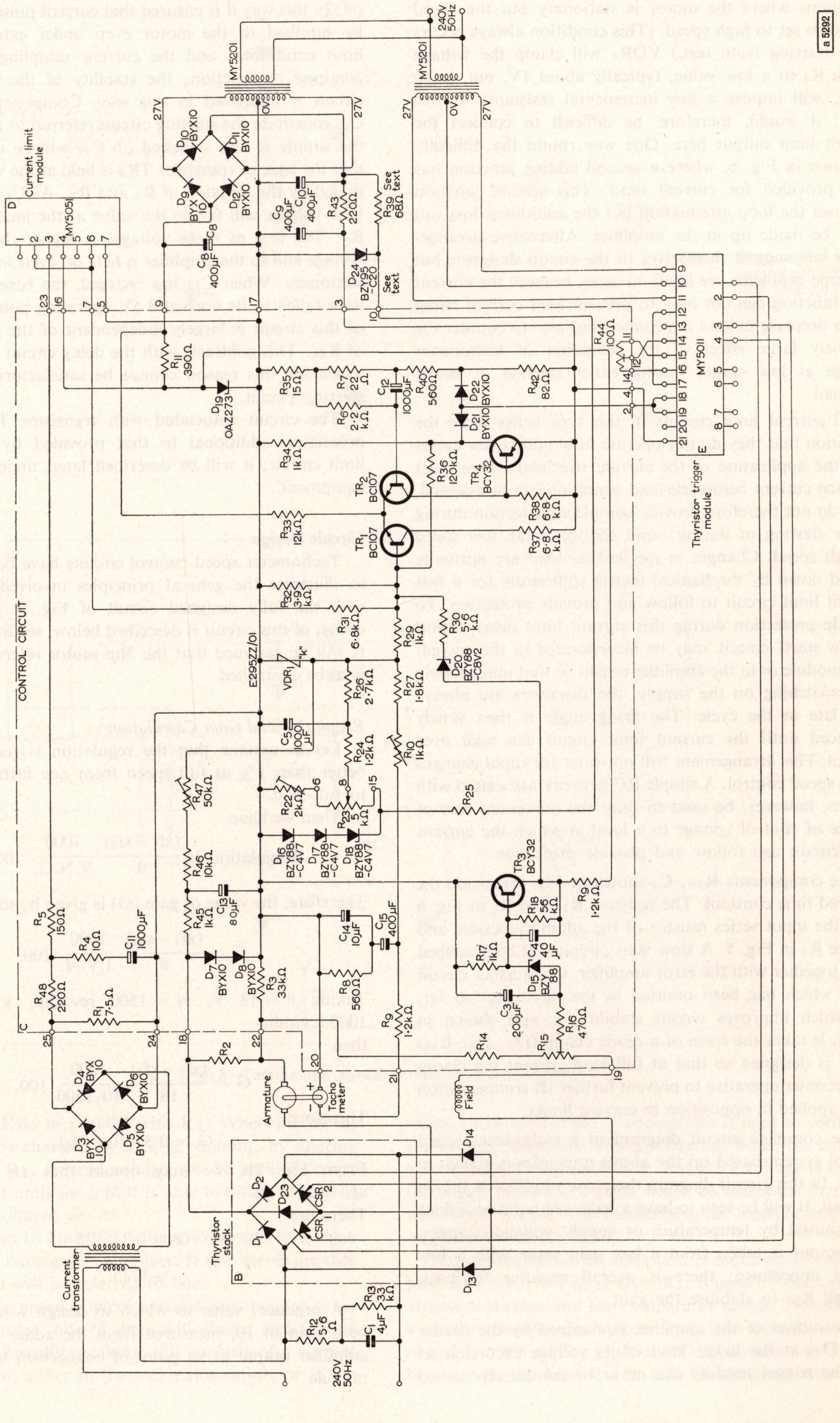

Fig. 7—Practical tachometer speed control circuit

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pages 82 *and* 87, Figs. 7 *and* 11

The asymmetrical voltage dependent resistor VDR_1 should have its cathode (marked 'K') connected to the line joining terminal 17 to R_3, and its anode connected to R_{47}, R_{26}, etc.

page 99, *right hand column*

In the instructions for setting resistors, the first instruction should read:

(i) Set R_{15} with its slider at the end nearer to terminal 19.

IR compensation

If, in addition to the regulation provided by the tachometer circuit, it is required that the regulation should be maintained at 2% by IR compensation down to 1/20 of full speed, the compensation circuit can be designed as follows (the component numbers refer to Fig. 6).

At 1/20 of the full-speed, a = 0·026.

If T_r = 0·01, then xf_1 = [illegible] (from Fig. 2c),

and if T_r = 0·05, then xf_2 = [illegible] (from Fig. 2b).

Therefore,

$$(xf_1 - xf_2) = 19°$$

The control voltage fed into the firing circuit must change by 1×([illegible]/18), or 1·05V, to produce this change in firing angle. The change in output from the tachometer at 1/20 full speed for a change in speed of 2% is given by,

$$\delta V_{(tachometer)} = \frac{1500 \times [illegible]}{20 \times 1000 \times 100}$$

$$= 0·015V.$$

If the control loop gain = 10, then the output voltage change $= 10 \times 0·015$

$$= 0·15V.$$

The difference between the required voltage and the amplified tachometer output, namely (1·05−0·15) or 0·9V, must be provided by the IR compensation circuit.

If the control circuit gain at the IR compensation input were also 10, the required control signal across C_1 would be (0·9/10) or 0·09V.

At full load the motor armature current is 25A and, at one-fifth load, the current will be approximately 5A. Therefore the change in current through R_6 is 20A. If for the first approximation we neglect the drop across R_7, then the value of R_6 is given by:

$$R_6 = 0·09/20 = 0·0045\Omega.$$

It is often more convenient to make R_6 considerably larger than this. The attenuation into the adder can then be increased by making R_8 larger. This has two advantages. The impedance of the smoothing circuit R_7, C_1 can be increased so that the capacitor can be smaller, and since R_8 can now be several times greater than R_1 and R_2, the addition of IR compensation does not load the adder. The optimum value of compensation can be adjusted by altering R_8 without appreciable interaction with the remainder of the circuit.

For example, if R_6 is made 0·045Ω, then R_8 will be approximately $10 \times R_1$ and the attenuation of the adder will be almost unchanged by the addition of IR compensation.

If the form factor of the armature circuit is approximately 1·0, the maximum dissipation (P) of R_6 at full-load torque will be approximately given by the equation,

$$P \simeq 25^2 \times 0·045$$

$$\simeq 28·5W.$$

In practice, when the average armature current is 25A, the r.m.s. current will be somewhat larger. The power dissipation will be increased in proportion to the square of the form factor which may be read from Fig. 8. The dissipation increase is unlikely to exceed 40%. It is interesting to note that the IR compensation has progressively less effect on the performance as the speed is increased. For example at full speed and maximum torque the trigger module input voltage will have to be about 5V. The IR compensation circuit will provide 0·9V as before, but in this case it only represents about 20% of the total control signal, whereas at 1/20 speed it provided 85%. At very low speeds the control signal from the IR compensation circuit approaches 100% of the whole.

Adder and Feedback Loop

Before proceeding further with the design of the complete control circuit the complex adder and error amplifier circuit will be analysed in detail. A complete design example is compared with a practical working circuit. As a first step in simplifying the circuit all capacitors are omitted, to give the steady state condition, and all series combinations of resistors, including internal resistance of tachometer, are combined to give the basic resistive network shown in Fig. 9a. We are only concerned in this analysis with the effect on the output voltage of changes in the various input voltages to the adder network. These steady state voltages are not of interest and the effective loop gain from any one input can be calculated while assuming that all other inputs remain constant. All input voltages with the exception of the one concerned, can therefore be considered to be short circuited. In addition, all parallel combinations of resistors can be combined to give a single equivalent component. For the tachometer input, the equivalent circuit can then be reduced to that of Fig. 9b.

The amplifier input network can be further simplified, by Thévenin's Theorem, into an equivalent simple generator and series resistance. The overall loop gain can then be calculated using the usual feedback equations, and taking into account the input attenuation.

From Fig. 9b, the equivalent resistance between points A and B is given by:

$$R_{AB} = \frac{R_c\left[R_6 + \dfrac{R_a . R_b}{(R_a+R_b)}\right]}{R_c + R_6 + \dfrac{R_a . R_b}{(R_a+R_b)}}.$$

The input voltage V_1 appears across terminals CD attenuated. Therefore,

$$V_{CD} = V_1 . \frac{R_b(R_6+R_c)}{R_a(R_b+R_6+R_c)+R_b(R_6+R_c)}.$$

This voltage is further attenuated before it reaches the amplifier input, where,

$$V_{AB} = V_{CD} . \frac{R_c}{R_6+R_c},$$

or

$$V_{AB} = V_1 . \frac{R_b(R_c+R_6)}{R_a(R_b+R_c+R_6)+R_b(R_c+R_6)} . \frac{R_c}{(R_c+R_6)}.$$

The equivalent circuit then becomes as shown in Fig. 9c. From feedback amplifier theory:

$$\text{Gain } \frac{V_o}{V_{AB}} = \frac{A}{1+\beta A},$$

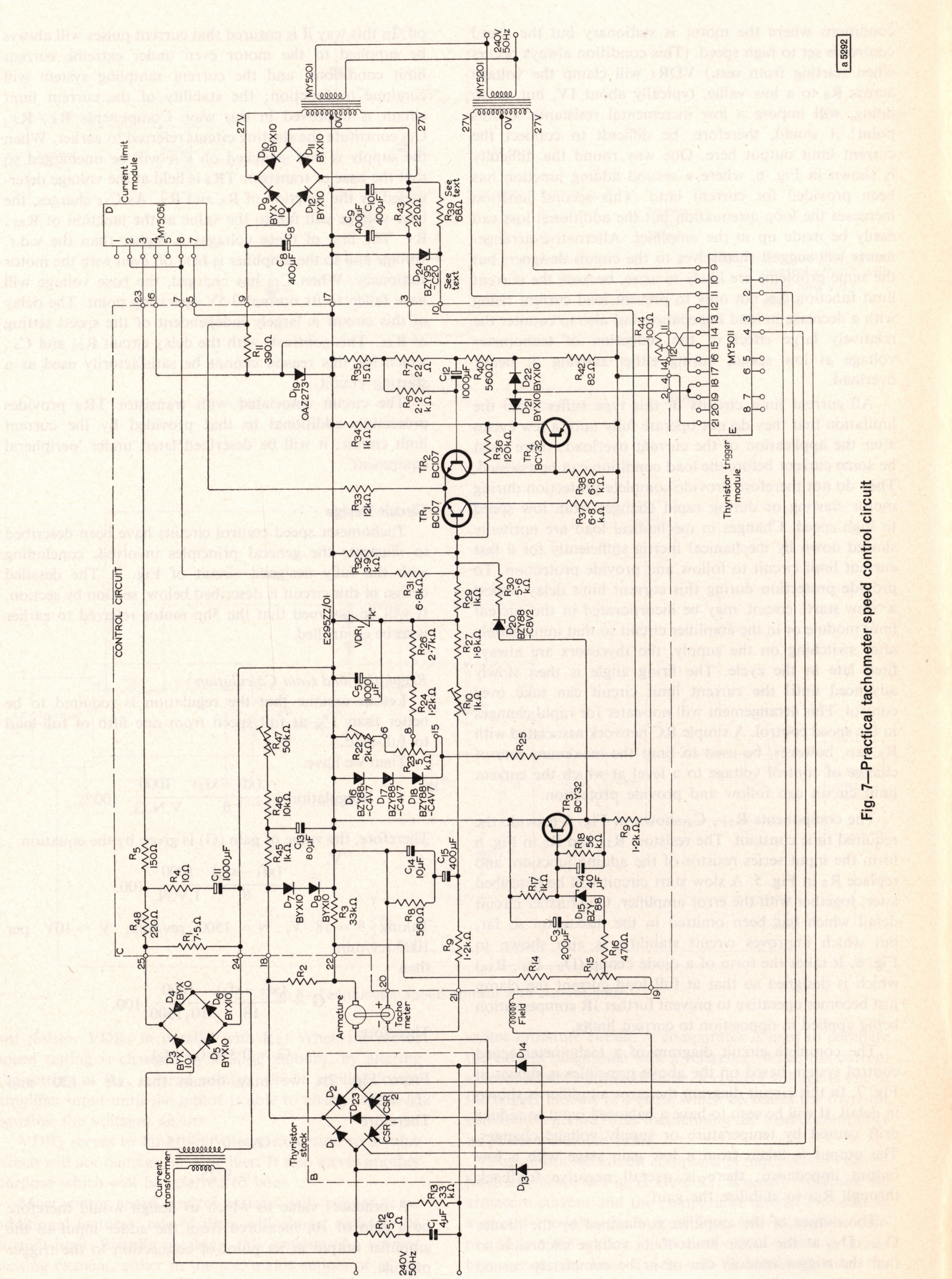

Fig. 7—Practical tachometer speed control circuit

IR compensation

If, in addition to the regulation provided by the tachometer circuit, it is required that the regulation should be maintained at 2% by IR compensation down to 1/20 of full speed, the compensation circuit can be designed as follows (the component numbers refer to Fig. 6).

At 1/20 of the full-speed, $a = 0{\cdot}026$.

If $T_f = 0{\cdot}01$, then $xf_1 = 166°$ (from Fig. 2c),

and if $T_f = 0{\cdot}05$, then $xf_2 = 147°$ (from Fig. 2b). Therefore,

$$(xf_1 - xf_2) = 19°.$$

The control voltage fed into the firing circuit must change by $1\times(19/18)$, or 1·05V, to produce this change in firing angle. The change in output from the tachometer at 1/20 full speed for a change in speed of 2% is given by,

$$\delta V_{(\text{tachometer})} = \frac{1500\times10\times2}{20\times1000\times100}$$

$$= 0{\cdot}015\text{V}.$$

If the control loop gain = 10, then the output voltage change $= 10\times0{\cdot}015$

$$= 0{\cdot}15\text{V}.$$

The difference between the required voltage and the amplified tachometer output, namely $(1{\cdot}05 - 0{\cdot}15)$ or 0·9V, must be provided by the IR compensation circuit.

If the control circuit gain at the IR compensation input were also 10, the required control signal across C_1 would be $(0{\cdot}9/10)$ or 0·09V.

At full load the motor armature current is 25A and, at one-fifth load, the current will be approximately 5A. Therefore the change in current through R_6 is 20A. If for the first approximation we neglect the drop across R_7, then the value of R_6 is given by:

$$R_6 = 0{\cdot}09/20 = 0{\cdot}0045\Omega.$$

It is often more convenient to make R_6 considerably larger than this. The attenuation into the adder can then be increased by making R_8 larger. This has two advantages. The impedance of the smoothing circuit R_7, C_1 can be increased so that the capacitor can be smaller, and since R_8 can now be several times greater than R_1 and R_2, the addition of IR compensation does not load the adder. The optimum value of compensation can be adjusted by altering R_8 without appreciable interaction with the remainder of the circuit.

For example, if R_6 is made 0·045Ω, then R_8 will be approximately $10\times R_1$ and the attenuation of the adder will be almost unchanged by the addition of IR compensation.

If the form factor of the armature circuit is approximately 1·0, the maximum dissipation (P) of R_6 at full-load torque will be approximately given by the equation,

$$P \simeq 25^2\times0{\cdot}045$$

$$\simeq 28{\cdot}5\text{W}.$$

In practice, when the average armature current is 25A, the r.m.s. current will be somewhat larger. The power dissipation will be increased in proportion to the square of the form factor which may be read from Fig. 8. The dissipation increase is unlikely to exceed 40%. It is interesting to note that the IR compensation has progressively less effect on the performance as the speed is increased. For example at full speed and maximum torque the trigger module input voltage will have to be about 5V. The IR compensation circuit will provide 0·9V as before, but in this case it only represents about 20% of the total control signal, whereas at 1/20 speed it provided 85%. At very low speeds the control signal from the IR compensation circuit approaches 100% of the whole.

Adder and Feedback Loop

Before proceeding further with the design of the complete control circuit the complex adder and error amplifier circuit will be analysed in detail. A complete design example is compared with a practical working circuit. As a first step in simplifying the circuit all capacitors are omitted, to give the steady state condition, and all series combinations of resistors, including internal resistance of tachometer, are combined to give the basic resistive network shown in Fig. 9a. We are only concerned in this analysis with the effect on the output voltage of changes in the various input voltages to the adder network. These steady state voltages are not of interest and the effective loop gain from any one input can be calculated while assuming that all other inputs remain constant.

All input voltages with the exception of the one concerned, can therefore be considered to be short circuited. In addition, all parallel combinations of resistors can be combined to give a single equivalent component. For the tachometer input, the equivalent circuit can then be reduced to that of Fig. 9b.

The amplifier input network can be further simplified, by Thévenin's Theorem, into an equivalent simple generator and series resistance. The overall loop gain can then be calculated using the usual feedback equations, and taking into account the input attenuation.

From Fig. 9b, the equivalent resistance between points A and B is given by:

$$R_{AB} = \frac{R_c\left[R_6 + \dfrac{R_a.R_b}{(R_a+R_b)}\right]}{R_c + R_6 + \dfrac{R_a.R_b}{(R_a+R_b)}}.$$

The input voltage V_1 appears across terminals CD attenuated. Therefore,

$$V_{CD} = V_1.\frac{R_b(R_6+R_c)}{R_a(R_b+R_6+R_c)+R_b(R_6+R_c)}.$$

This voltage is further attenuated before it reaches the amplifier input, where,

$$V_{AB} = V_{CD}.\frac{R_c}{R_6+R_c},$$

or

$$V_{AB} = V_1.\frac{R_b(R_c+R_6)}{R_a(R_b+R_c+R_6)+R_b(R_c+R_6)}.\frac{R_c}{(R_c+R_6)}.$$

The equivalent circuit then becomes as shown in Fig. 9c. From feedback amplifier theory:

$$\text{Gain}\ \frac{V_o}{V_{AB}} = \frac{A}{1+\beta A},$$

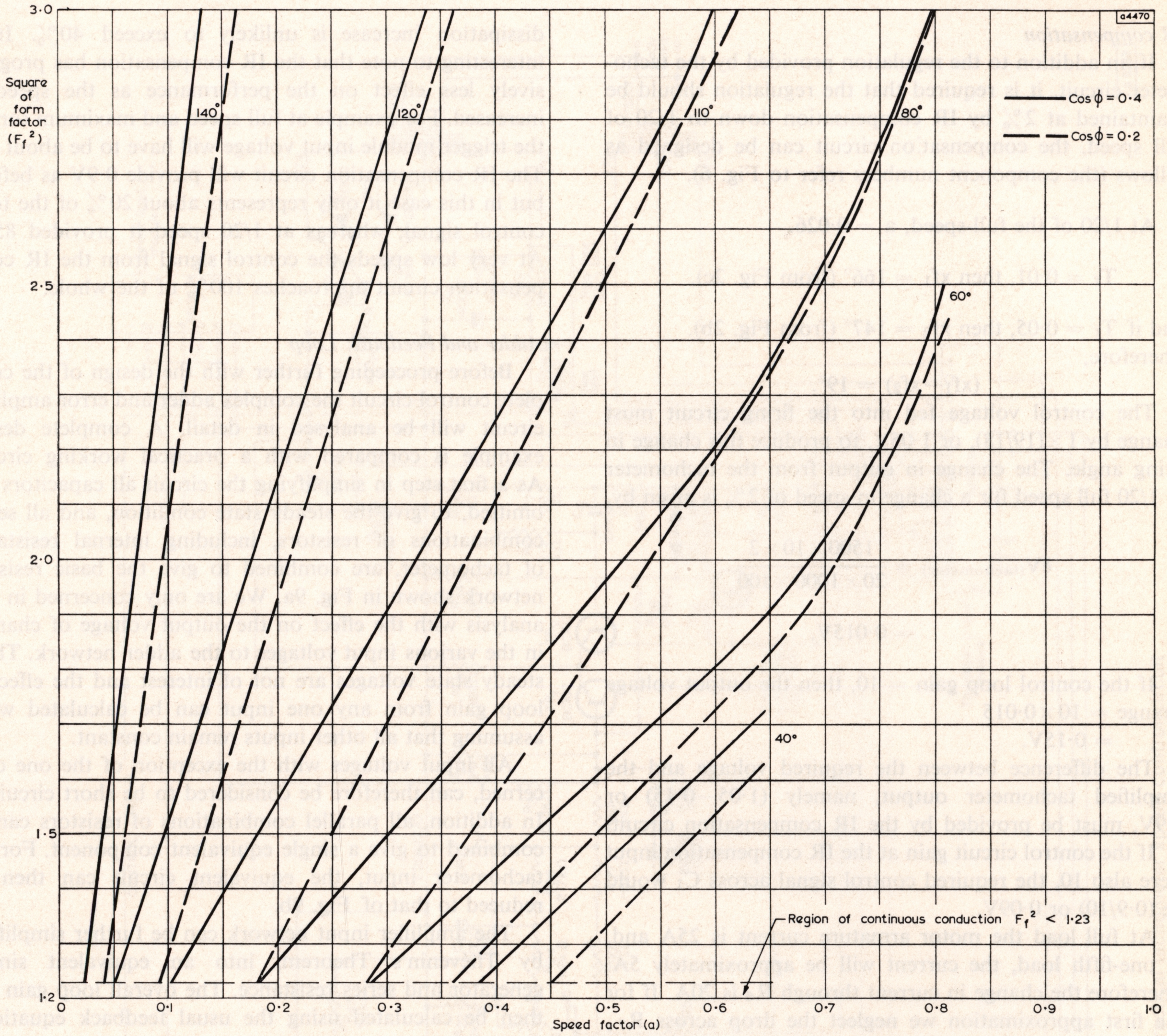

Fig. 8—Curves for derivation of square of form-factor, using speed factor, firing angle and cos φ

where

$$\beta = \frac{R_{AB}}{R_{10}+R_{AB}}.$$

Therefore

$$\beta = \frac{1}{1+(R_{10}/R_{AB})}.$$

The required overall circuit gain (G) is given by

$$G = V_o/V_1.$$

Therefore

$$G = \frac{V_o}{V_{AB}} \cdot \frac{V_{AB}}{V_1}.$$

Therefore

$$G = \frac{A}{(1+\beta A)} \cdot \frac{R_b(R_c+R_6)}{R_a(R_b+R_c+R_6)+R_b(R_c+R_6)} \cdot \frac{R_c}{(R_c+R_6)}$$

where

$$\beta = \frac{1}{1+(R_{10}/R_{AB})}$$

and

$$R_{AB} = \frac{R_c[R_6(R_a+R_b)+R_a . R_b]}{(R_a+R_b)(R_c+R_6)+R_a . R_b}.$$

If the actual component values for the circuit of Fig. 7 are inserted into the above equations we find that,

$$R_a \simeq 4·2\text{k}\Omega,$$
$$R_b \simeq 3·9\text{k}\Omega,$$
$$R_6 \simeq 1·0\text{k}\Omega,$$
$$R_c \simeq 2·5\text{k}\Omega,$$
$$R_{10} \simeq 120\text{k}\Omega.$$

Using these equivalent component values in the above equations gives,

$$G = \frac{1}{(1+\beta A)} \times 0·216$$

and

$$\beta = 0·0113.$$

Taking A as 100 gives,

$$G = \frac{100}{(1+1·13)} \times 0·216$$
$$= 10·1.$$

This may be compared with a measured gain of 10·7 in a particular circuit, confirming that the analytical approach agrees well with practical measurements. The above calculation yields the overall loop gain when the control circuit error signal is received from the tachometer, that is, it is applicable when we are considering the control circuit for stabilising motor speed. Similarly the loop gain for the IR compensation circuit input is about 1·2.

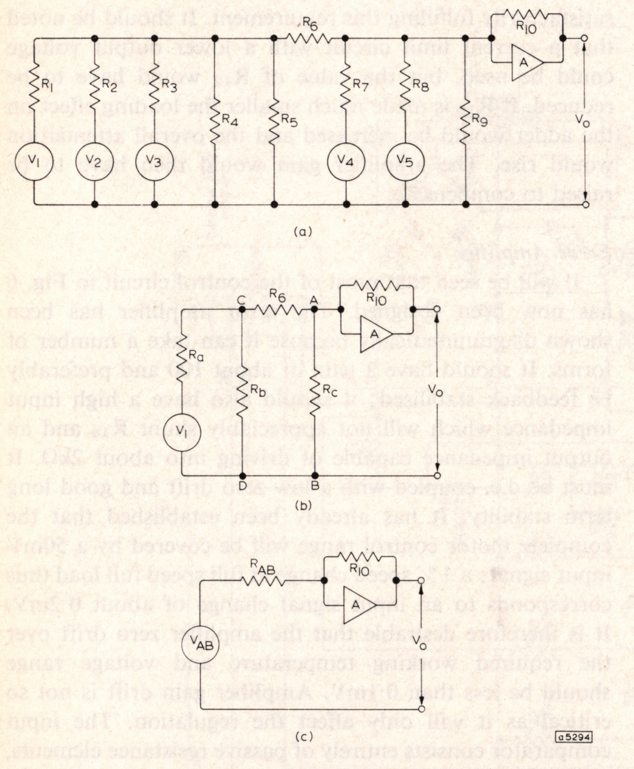

Fig. 9—(a) Resistor adder and amplified equivalent circuit
(b) Simplified equivalent circuit
(c) Basic circuit

Under starting or severe overload conditions the instantaneous error voltage, appearing at the adder output, will be large. When this voltage exceeds about 1V, the v.d.r. will appear in the circuit as a low resistance and considerably attenuate the error signal. Under these conditions, to a first approximation, the v.d.r. may be considered as a short circuit so that the first part of the adder can be omitted from the analysis. The equivalent circuit of Fig. 9b with $R_b = 0$ can then be redrawn with R_6 in parallel with R_c.

The gain equations from either of the inputs to the second adding junction can then be derived in a similar way to that described for the first junction.

Predicted and Measured Performance

The circuit design has now reached a point where it is possible to calculate the regulation to be expected from the circuit, and to compare this with actual measured performance.

From the earlier discussion it will be seen that a tachometer loop gain of 10 gave a calculated regulation at maximum speed of 0·9% for a load change from 1/5 load to full-load. Examination of the firing angle curves will show that at very small loads the speed will increase considerably, and therefore the actual no-load to full-load regulation will depend critically on the no-load windage and frictional losses of the actual machine. Since true no-load speed is of little practical interest a minimum load of 1/5 of full-load has been chosen for these calculations.

We will now consider the effect of IR compensation. The voltage drop across the IR compensation resistor R_6 (Fig. 6) will change by an amount given by,

$$\delta V_I = R_6(I_{max} - I_{min}),$$

where

$$I = \text{Armature current.}$$

This voltage change will appear at the input to the firing circuit after amplification by the feedback amplifier. There will also be an input to the amplifier from the tachometer. Let this voltage be δV_T for a certain small speed change. This voltage change will also be amplified before it appears at the input to the firing circuit where it will add to the IR compensation signal. If the gain for the IR compensation loop is given by G_I and for the tachometer loop by G_T, then the above statement can be expressed as follows:

$$\delta V_I . G_I + \delta V_T . G_T = V_c$$

where V_c is the control voltage fed to the trigger circuit.

Now if V_T is the tachometer output voltage, then

$$\text{Regulation} = \frac{\delta V_T}{V_T} \times 100\%.$$

Combining these two equations gives,

$$\text{Regulation} = \frac{V_c - \delta V_I . G_I}{V_T . G_T} . 100\%.$$

If we now substitute practical values into this design equation we can compare the theoretical value obtained with the value obtained by measurements.

Taking $R_6 = 0·036\Omega$, $I_{max} = 25\text{A}$ and $I_{min} = 5\text{A}$, then

$$\delta V_I = 0·72\text{V}$$
$$G_I = 1·2.$$

Therefore

$$\delta V_I . G_I = 0·86\text{V}.$$

At 1/20 of full speed a tachometer with sensitivity 10V per 1000 rev/min will give $V_T = 0·75\text{V}$. We also have $G_T = 10$.

Therefore

$$V_T . G_T = 7·5\text{V}.$$

If regulation is good, the value of V_c will be very near to that required to maintain constant speed over the load range. In this example, $V_c = 1·05\text{V}$.

We can therefore estimate the system regulation.

$$\text{Regulation} = \frac{1·05 - 0·86}{7·5} . 100\%$$
$$= 2·5\%.$$

The overall system performance can now be summarised as follows. The calculated percentage regulation figures are for a load change of 1/5 of full load to full load; measured regulation results are shown in brackets.

Full-speed regulation = 0·9% (0·9%)
1/3 full-speed regulation = 2·4% (2·5%)
1/20 full-speed regulation = 2·5% (3·0%)

It will be seen that the agreement between the theoretical and measured regulation is very satisfactory; the agreement is as much limited by the difficulty of measurement as by the accuracy of calculation; thus these results are given to no more than two significant figures. Regulation at other speeds can be obtained from Fig. 10.

Current Limit

The current limit characteristics will depend upon the method which is used to sense the load current. It is possible to obtain a control signal which is proportional to the average, or r.m.s., or peak value of either the supply line current or the motor armature current. All will give quite widely differing results and the choice will depend to some extent upon the application. This aspect of the circuit design will be dealt with later. It will suffice at present to consider the output of the current limit module and its integration into the remainder of the control circuit.

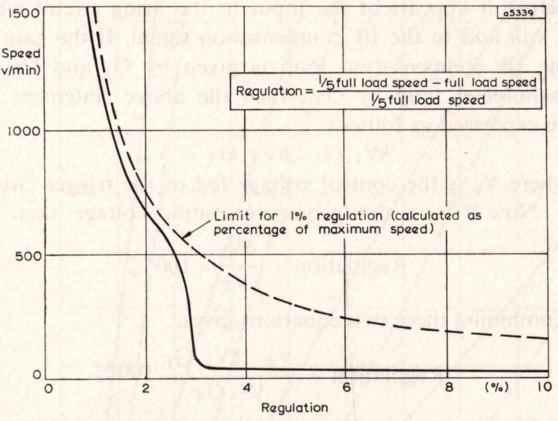

Fig. 10—Regulation versus speed for tachometer speed control circuit

The worst case which the current limit circuit will have to protect against is a stalled motor with the speed control set to maximum speed. Under these conditions VDR$_1$ will have a low resistance and the voltage across it will be about 1V. (VDR$_1$ is the E295ZZ/01). The trigger module (Mullard MY5011) has a control sensitivity of 18°/volt; the maximum control voltage required for motor speed control will therefore be about 5V for a firing angle range from 70° to 160°. With an amplifier gain of 100, this requires an input signal to the amplifier of only 50mV. If the amplifier is biased to a mid operating point of about 0·5V, therefore, there will be no risk of driving it outside the linear region. The maximum input, however, will be limited by the v.d.r. to 1V. Under these conditions the maximum input signal which the amplifier will ever receive will be 0·5V above the nominal operating point. At the amplifier output this will be $(0{\cdot}5 \times 100) = 50\text{V}$.

The current limit loop gain is 9 for this particular circuit. Therefore the minimum current limit input signal which will just counter this maximum input signal is $50/9 = 5{\cdot}5\text{V}$ at terminal D. It is therefore a fundamental design requirement that the current limit output voltage must be capable of exceeding in this case 5·5V. The current limit module used (MY5051) gives an output of at least 6V, satisfactorily fulfilling this requrement. It should be noted that a current limit circuit with a lower output voltage could be used, but the value of R_{12} would have to be reduced. If R_{12} is made much smaller the loading effect on the adder would be increased and the overall attenuation would rise. The amplifier gain would then have to be raised to compensate.

Error Amplifier

It will be seen that most of the control circuit in Fig. 6 has now been designed. The error amplifier has been shown diagrammatically because it can take a number of forms. It should have a gain of about 100 and preferably be feedback stabilised; it should also have a high input impedance which will not appreciably shunt R_{14} and an output impedance capable of driving into about 2kΩ. It must be d.c. coupled with a low zero drift and good long term stability. It has already been established that the complete motor control range will be covered by a 50mV input signal; a 1% speed change at full speed full load thus corresponds to an input signal change of about 0·2mV. It is therefore desirable that the amplifier zero drift over the required working temperature and voltage range should be less than 0·1mV. Amplifier gain drift is not so critical as it will only affect the regulation. The input comparator consists entirely of passive resistance elements, and so it will be largely free from drift effects. An amplifier which has proved to be satisfactory is shown in the full circuit diagram Fig. 7. The performance of this amplifier is as follows:

Open loop voltage gain = 100
Input impedance > 25kΩ
Output impedance < 1kΩ.

When used in conjunction with the motor control circuit, with a feedback resistor $R_{36} = 56\text{k}\Omega$, the typical overall performance is as follows:

Closed loop gain = 7·5
Input level change = 0·5mV for supply change of 200V to 250V
Input level change = 3·5μV/degC ambient change (measured at the amplifier input)

Higher amplifier loop gain is possible. For example, with $R_{36} = 100\text{k}\Omega$, Gain = 10, and for $R_{36} = 220\text{k}\Omega$, Gain = 14, but at increased gain the overall amplifier stability is reduced to some extent.

ARMATURE VOLTAGE SPEED CONTROL

The tachometer control circuit, which has been described above, made use of a motor mounted d.c. tachometer to measure the motor speed and provide the required feedback control signal. The circuit in Fig. 11, to be described, makes use of a bridge to measure the motor generated back e.m.f.; this is proportional to motor speed at constant field. The control circuit is very similar to that used with the tachometer, and consequently only the points which differ will be described in detail.

Circuit Description

The armature voltage bridge consists of a resistance in series with the motor armature R_2 to form one bridge arm; the other arm consists of R_{20} and R_{21}. The inductance

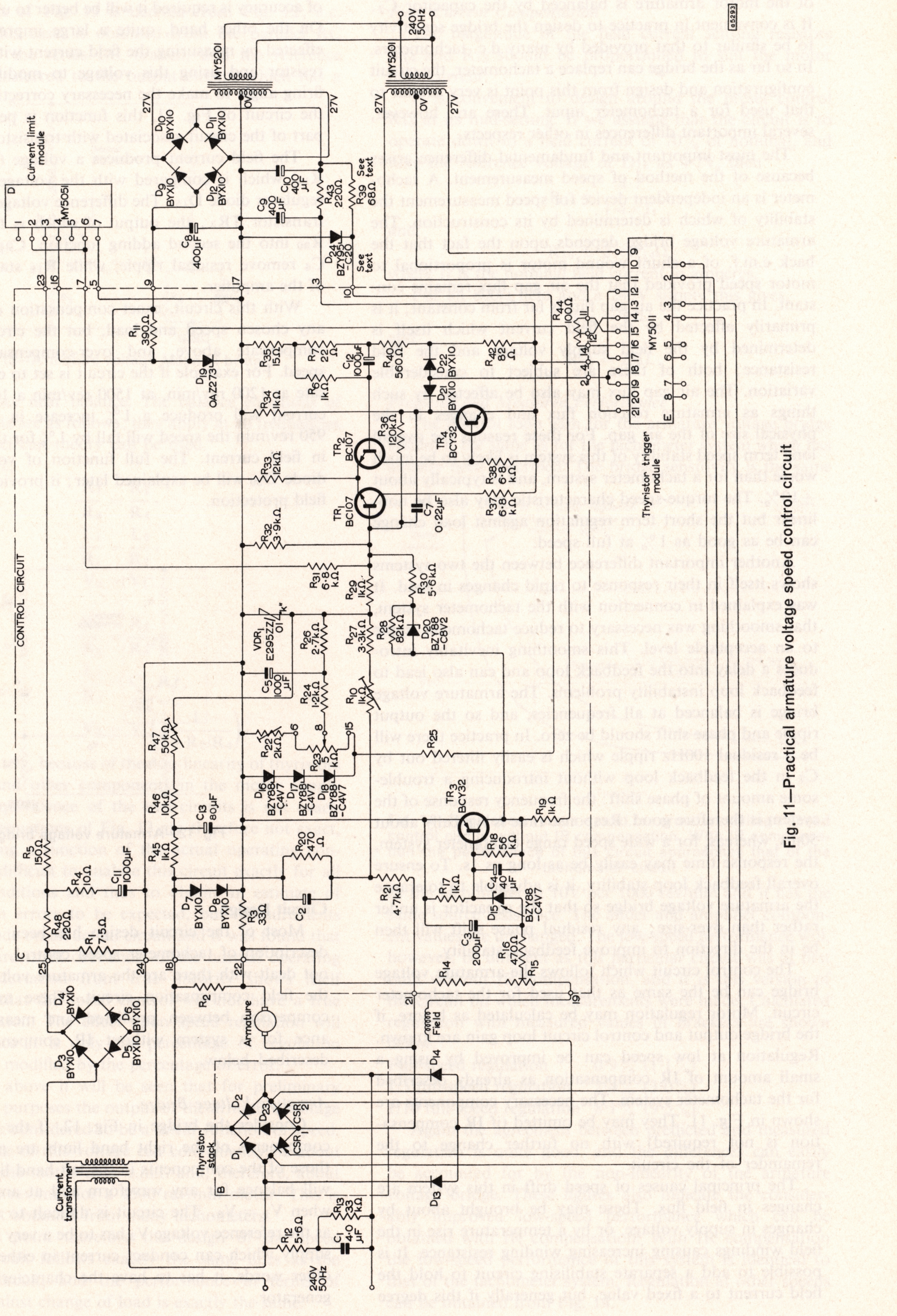
a5333
240V 50Hz
MY5201
27V
0V
27V
Current limit module
MY5051
D
BYX10
C_{10} 40µF
C_8 400µF
C_9 40µF
R_{43} 220Ω
R_{39} See text
68Ω See text
D_{24} BZY88-C20
See text
MY5011
Thyristor trigger module
R_{44} 100Ω
R_{11} 390Ω
D_{19} OAZ213
R_{35} 15kΩ
R_{34} 1kΩ
R_{40} 560Ω
R_{42} 82Ω
C_{12} 1000µF
D_{22} BYX10
D_{21} BYX10
TR2 BC107
TR4 BCY32
TR1 BC107
R_{36} 120kΩ
R_{33} 12kΩ
R_{37} 6·8kΩ
R_{38} 6·8kΩ
C_7 0·22µF
R_{32} 3·9kΩ
R_{31} 6·8kΩ
R_{29} 1kΩ
R_{30} 5·6kΩ
R_{28} 82kΩ
D_{20} BZY88-C8V2
CONTROL CIRCUIT
VDR1 E295ZZ/01
C_5 100µF
R_{26} 2·7kΩ
R_{27} 3·3kΩ
R_{24} 12kΩ
R_{10} 1kΩ
R_{22} 2kΩ
R_{23} 1·5kΩ
R_{25}
R_{47} 50kΩ
D_{16} BZY88-C4V7
D_{17} BZY88-C4V7
D_{18} BZY88-C4V7
R_{46} 10kΩ
C_6 80µF
R_{20}
R_{21} 4·7kΩ
TR3 BCY32
R_{19} 1kΩ
R_{18} 56Ω
R_{17} 1kΩ
C_4 40µF
D_{15} BZY88-C4V7
C_3 200µF
R_{16} 470Ω
R_{15}
R_{14}
R_5 150Ω
R_4 10Ω
C_1 1000µF
R_{45} 1kΩ
R_{48} 220Ω
R_1 7·5Ω
D_7 BYX10
D_8 BYX10
R_3 33Ω
C_2
R_2
Armature
Field
D_4 BYX10
D_6 BYX10
D_3 BYX10
D_5 BYX10
D_{14}
D_{13}
D_2
D_{23}
CSR
D_1
Thyristor stack
Current transformer
R_{12} 3·6Ω
R_9 33kΩ
C_1 40µF
240V 50Hz
B
C
A
E

Fig. 11—Practical armature voltage speed control circuit

of the motor armature is balanced by the capacitor C_2. It is convenient in practice to design the bridge sensitivity to be similar to that provided by many d.c. tachometers. In so far as the bridge can replace a tachometer, the circuit configuration and design from this point is very similar to that used for a tachometer input. There are, however, several important differences in other respects.

The most important and fundamental difference arises because of the method of speed measurement. A tachometer is an independent device for speed measurement the stability of which is determined by its construction. The armature voltage bridge depends upon the fact that the back e.m.f. of a shunt wound motor is proportional to motor speed provided that the air gap flux remains constant. In practice the air gap flux is far from constant; it is primarily affected by the field current which itself is determined by the field supply voltage and the field resistance; both of these are subject to considerable variation. The air gap flux may also be affected by such things as armature reaction flux and changes in the physical size of the air gap. For these reasons the overall long term speed stability of this system is likely to be much worse than for a tachometer system, and is typically about ±10%. The torque-speed characteristic may also be non-linear but the short term regulation against load change can be as good as 1% at full speed.

Another important difference between the two systems shows itself in their response to rapid changes in load. It was explained in connection with the tachometer system, that smoothing was necessary to reduce tachometer ripple to an acceptable level. This smoothing inevitably introduces a delay into the feedback loop and can also lead to feedback loop instability problems. The armature voltage bridge is balanced at all frequencies, and so the output ripple and phase shift should be zero. In practice there will be a residual 100Hz ripple which is easily filtered out by C_7 in the feedback loop without introducing a troublesome amount of phase shift: the frequency response of the system is therefore good. Response time is typically about 50ms, whereas, for a wide speed range tachometer system, the response time may easily be as long as 1s. To ensure overall feedback loop stability, it is advisable to tolerance the armature voltage bridge so that the capacitor is under rather than over-size; any residual phase shift will then be in the direction to improve feedback stability.

The control circuit which follows the armature voltage bridge can be the same as that used for the tachometer circuit. Motor regulation may be calculated as before, if the bridge output and control circuit loop gain are known. Regulation at low speed can be improved by using a small amount of IR compensation as already described for the tachometer system. The necessary components are shown in Fig. 11. They may be omitted (if IR compensation is not required) with no further change to the remainder of the circuit.

The principal causes of speed drift in this system are changes in field flux. These may be brought about by changes in supply voltage, or by a temperature rise in the field windings causing increasing winding resistance. It is possible to add a separate stabilising circuit to hold the field current to a fixed value, but generally if this degree of accuracy is required it will be better to use a tachometer. On the other hand, quite a large improvement can be effected by measuring the field current with a small series resistor and using this voltage to modify the thyristor firing angle to make the necessary correction to speed. In the circuit of Fig. 11, this function is performed by the part of the circuit associated with transistor TR_3.

The field current produces a voltage (across R_{14} and R_{15}) which is compared with the voltage across voltage regulator diode D_{15}. The difference voltage is amplified by transistor TR_3, the output of which is fed via R_{28} and R_{30} into the second adding junction. Capacitors C_3 and C_4 remove residual ripple, while R_{18} stabilises the gain of the amplifier.

With this circuit, exact compensation can be made at any chosen speed and load, but the circuit will under-compensate above, and over-compensate below this speed. For example if the circuit is set to exactly compensate at 1200 rev/min, at 1500 rev/min a 10% fall in field current will produce a 1% increase in speed while at 950 rev/min the speed will fall by 1% for the same change in field current. The full function of voltage regulator diode D_{20} will be explained later; it provides open circuit field protection.

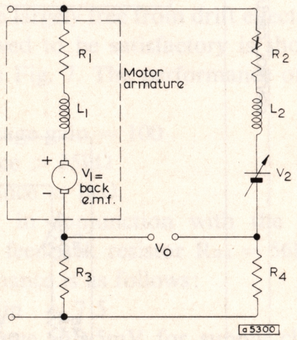

Fig. 12—Armature voltage bridge

Circuit Design

Most of the circuit design has been covered in the description of tachometer speed control. The only parts not dealt with there are the armature voltage bridge and the field compensation circuit. These together with a comparison between predicted and measured performance for a system without IR compensation will be described below.

Armature Voltage Bridge

Consider the bridge in Fig. 12. If the impedances of components of the right hand limb are proportional to those of the components in the left hand limb, the bridge will balance for any waveform and at any motor speed when $V_1 = V_2$. The circuit is difficult to use in practice as the reference voltage V_2 has to be a very low impedance supply which can conduct current in either direction. In other words, it has to have the characteristics of a d.c. generator.

The circuit of Fig. 13 is derived from Fig. 12; a.c. balance, in this case, is obtained from the parallel RC equivalent of the series LR branch, and the reference voltage is transferred to the output circuit, which has a high impedance near balance. The balance equations of the two circuits are as follows; they apply at all frequencies where linear circuit theory can be used.

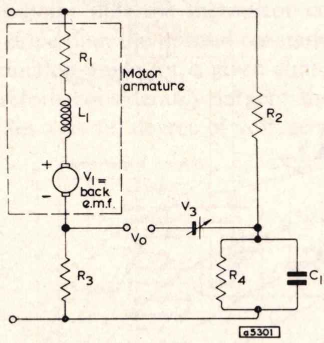

Fig. 13—Armature voltage Maxwell bridge

For Fig. 12,

$$\frac{R_1}{R_3} = \frac{R_2}{R_4},$$

$$\frac{L_1}{R_1} = \frac{L_2}{R_2},$$

$$V_2 = V_1.$$

For Fig. 13,

$$\frac{R_1}{R_3} = \frac{R_2}{R_4},$$

$$\frac{L_1}{R_1} = C_1 . R_4,$$

$$V_3 = V_1\left(\frac{1}{1+R_2/R_4}\right).$$

Unfortunately, because of the non-linearity of thyristors and diodes and other components in the motor supply circuit, the equivalence of the two circuits is not exact. The balance equations of Fig. 13 are therefore not exact, the error being a function of the actual operating conditions. It is difficult to analyse this circuit exactly for all operating conditions and thus to obtain an estimate of the maximum errors to be expected, but measurements have been made on an actual circuit and it was found that these errors are small over most of the normal operating range. The deviation from true linearity shows itself primarily in that the speed control will not have a true linear calibration. The effect upon speed regulation will be small, since it is only the percentage of regulation which will be modified by the percentage of error.

From the above it will be seen that for preliminary circuit design purposes the output of the armature voltage bridge may be considered to be a linear function of speed. The magnitude of this voltage can conveniently be about one tenth of the motor back e.m.f. and, therefore, about 10V per 1000 rev/min for a 1500 rev/min motor. This is similar to that obtained from many tachometers.

From this point therefore the design of the speed stabilising circuit follows exactly the same lines as the tachometer circuit, and the performance in terms of regulation against change of load is exactly the same.

Field Current Compensation

The magnitude of the field current sensing resistors R_{14} and R_{15} should be proportioned to suit the motor field current.

It is convenient to design around the preset control R_{15} at its mid-setting. If we assume that the circuit is to operate down to a field current of 70% of nominal, and take the reference voltage of D_{15} as 4·7V, then

$$R_{14} \simeq \frac{4{\cdot}7}{1{\cdot}4 \times I_f}\Omega,$$

and,

$$R_{15} \simeq 2 \times R_{14}\Omega,$$

where I_f is the nominal field current in amperes.

For an 800Ω field and a 240V supply this gives,

$$R_{14} = 12\Omega$$

and

$$R_{15} = 25\Omega.$$

Predicted and Measured Performance

The control loop gain for the armature voltage bridge system is the same as for the tachometer system which has already been described. The bridge sensitivity is the same as the tachometer at 10V per 1000 rev/min and it has the same source resistance. The adder and amplifier component values are unchanged. The low-speed performance will, however, be different as this particular circuit was at first built without IR compensation, and it is this version of the circuit on which measurements were made. Calculated regulation with measured values in brackets are shown below.

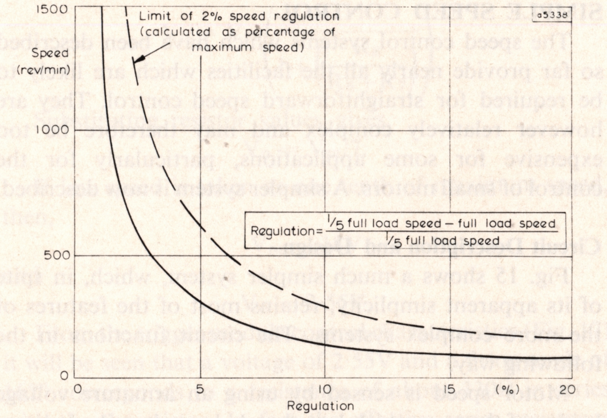

Fig. 14—Regulation versus speed for armature voltage speed control circuit without IR compensation. With IR compensation the performance of this circuit approaches that of the tachometer circuit

Full-speed regulation	0·9% (1·1%)
1/3 full-speed regulation	2·4% (3·0%)
1/10 full-speed regulation	8·1% (9·5%).

Again the agreement between predicted and measured regulation is quite good, and the differences can easily be accounted for by the non-linearity of the armature voltage bridge. These figures also indicate the considerably improved low-speed performance which can be obtained with IR compensation. With IR compensation the low-speed performance of this system is identical to that of the tachometer system. Regulation at other speeds can be obtained from Fig. 14.

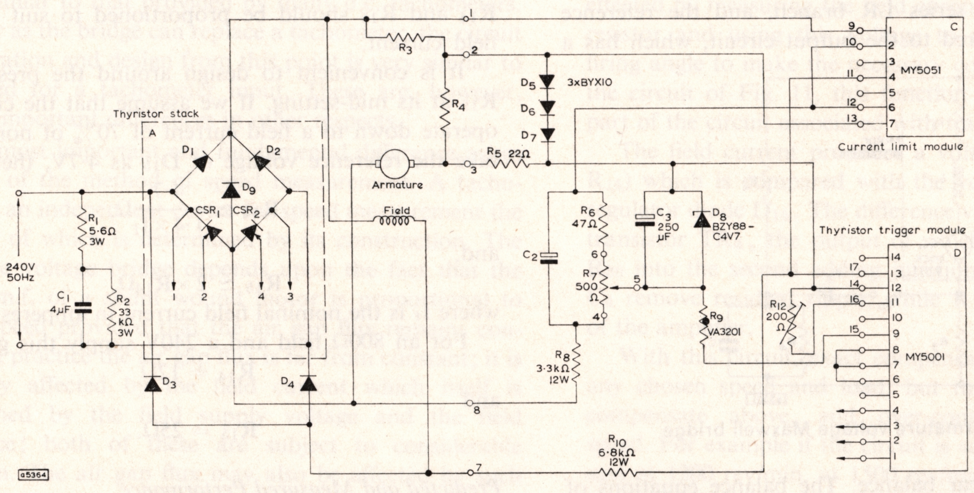

Fig. 15—Practical simple speed control circuit

SIMPLE SPEED CONTROL

The speed control systems which have been described so far provide nearly all the facilities which are likely to be required for straightforward speed control. They are however relatively complex and may therefore be too expensive for some applications, particularly for the control of small motors. A simpler system is now described.

Circuit Description and Design

Fig. 15 shows a much simpler system, which, in spite of its apparent simplicity, retains most of the features of the more complex systems. The circuit functions in the following way.

Motor speed is sensed by using an armature voltage bridge of the type which has already been described above. In this case, the bridge arms consists of R_4, R_8, and the network C_2, R_6, R_7. The bridge output is taken from the junction of the motor armature and R_4 and from the adjustable tap of the speed control resistor R_7. The output voltage is fed directly to the input terminals of the trigger module type MY5001 via a bead thermistor R_9, VA3201.

When the tapping point of R_7 is near the lower end, the motor speed will be low, because the maximum output voltage will be fed to the trigger module. As the tapping point is moved up, the motor speed will increase to produce a larger voltage to re-establish equilibrium within the system. The diode clamp (D_5, D_6, D_7 and R_5) prevents the feedback component, due to armature current, from increasing indefinitely under overload conditions. The clamp is designed to operate at just over full-load current, at the point where the current limit circuit begins to be effective.

Capacitor C_3 provides some final smoothing to remove residual ripple from the bridge output.

Voltage regulator diode D_8 prevents the bridge output voltage from ever exceeding about 5V. This could otherwise arise if the speed control is quickly set to the low-speed setting, when the motor is rotating at high-speed and thus generating a high back e.m.f. An excessive voltage at this point could damage the trigger module. Limiting the voltage excursion in this way also permits C_3 to have a low working voltage. Thermistor R_9 serves three purposes: it ensures that the circuit is always set to a large trigger angle when switching on, so that large starting current surges are avoided; it serves a similar function when the motor speed is rapidly changed from a low to a high value; it increases the feedback loop gain by a factor of about 2. This latter function will be expanded upon when the circuit is analysed in more detail later.

Of the remaining components, R_3 provides an input voltage proportional to armature current for the current limit circuit, and R_{10} is the series voltage dropper resistor for the MY5001 power supply. The input transient suppression network is common to all systems and will be described separately.

Armature Voltage Bridge. Regulation

The control characteristics between terminals 9 and 10 of a typical MY5001 trigger module are shown by the full lines of Fig. 16 for a range of series resistance values and control voltages. This family of curves can be moved to the right of the diagram and slightly modified in shape by introducing additional resistance between terminals 11 and 12. The characteristics with 180Ω between terminals 11 and 12 are shown in Fig. 17. In the circuit in Fig. 15, the series resistance consists of the thermistor R_9 and the armature voltage bridge source resistance.

The control voltage is the bridge output voltage, the magnitude of which will depend upon motor speed and the setting of the speed control potentiometer R_7.

The thermistor R_9 is heated by the current passing through it; this current is a function of the firing angle delay of the MY5001 trigger module. The current into terminal 9 is large when the firing angle is small and very much smaller when the angle is large. The thermistor resistance is therefore variable and is a function of the control voltage. It is therefore convenient to draw another set of control characteristics including a thermistor in series with the control resistance. These characteristics are shown as broken lines in Figs. 16 and 17. It is interesting to note at this point that the thermistor control curves have a steeper slope than the normal resistance curves, the change in conduction angle for a given change in control voltage is therefore considerably larger; the thermistor, in effect, provides a useful degree of voltage amplification.

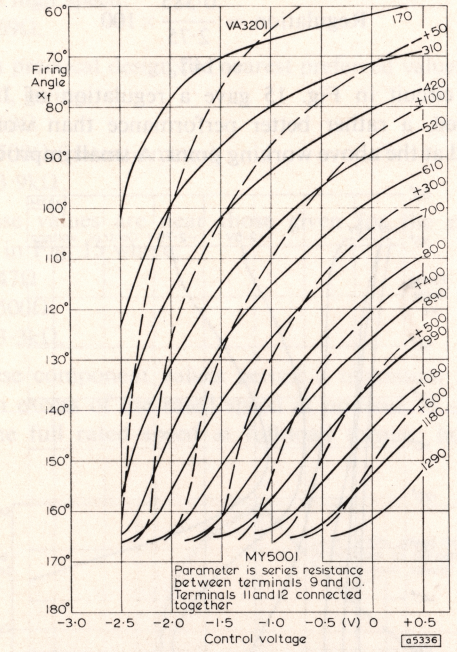

Fig. 16—Control characteristics for MY5001 module using various resistance values between terminals 9 and 10 with a shorting link between 11 and 12. (Dotted curves are for a VA3201 and resistance between terminals 9 and 10)

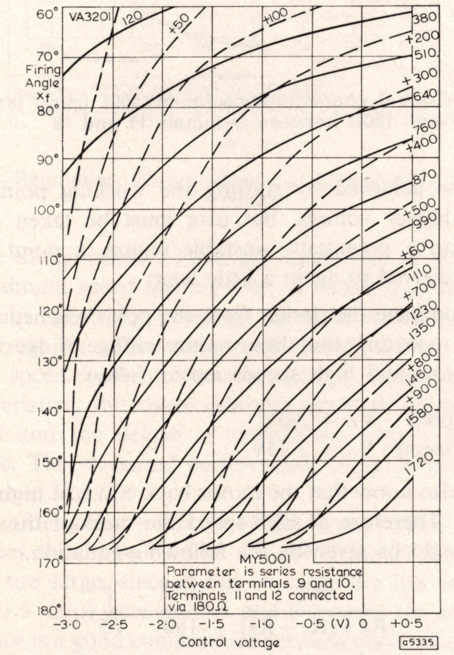

Fig. 17—Control characteristics for MY5001 module using various resistance values between terminals 9 and 10 with 180Ω between 11 and 12. (Dotted curves are for VA3201 and resistance between terminals 9 and 10)

Measurements show that the increase in control voltage sensitivity can lie between 1·5 to 1 and 2·5 to 1 when a series thermistor is used.

To illustrate the operation of this circuit in more detail we will consider the circuit of Fig. 15 with the component values given, and with the 5hp motor which was used as an example for the analysis of the previous control systems. This motor has a rated d.c. armature voltage of 180V. At full speed, the back e.m.f. will therefore be a little under 180 volts, and the d.c. speed factor may be taken to be 0·53. The full-load torque factor may be taken as 0·05 and $\cos\theta = 0{\cdot}4$. The essential motor data is therefore as follows.

Maximum speed $a = 0{\cdot}53$
Maximum torque $T_f = 0{\cdot}05$, therefore $x_f = 95°$ (Fig. 2b)
1/5 torque $T_f = 0{\cdot}01$, therefore $x_f = 120°$ (Fig. 2c)
1/10 speed $a = 0{\cdot}053$
Maximum torque $T_f = 0{\cdot}05$, therefore $x_f = 140°$ (Fig. 2b)
1/5 torque $T_f = 0{\cdot}01$, therefore $x_f = 160°$ (Fig. 2c)

When the speed control potentiometer R_7 is set to the low speed, the bridge output voltage (V_o) will be related to motor back e.m.f. (V_m) by the equation,

$$V_o = V_m\left(\frac{1}{1+\left(\frac{R_8}{R_6+R_7}\right)}\right).$$

Substituting resistor values gives,

$$V_o = V_m/7.$$

If low speed is taken as one tenth of maximum speed, then,

$$V_m = 18\text{V}$$

and

$$V_o = 2{\cdot}55\text{V}.$$

From the trigger module control characteristics (Fig. 17) it will be seen that a voltage of 2·55V and a firing angle of 162° corresponds to a resistance of about 330Ω in series with the thermistor. In fact the effective series resistance will be approximately 500Ω. This discrepancy can be accounted for in two ways; either the speed at the limit of the control setting is rather less than 1/10 full speed so that the back e.m.f. will be less than 2·55V, or the circuit tolerances are such that the voltage scale on the characteristics is effectively shifted. Either effect can be allowed for by a small shift in operating voltage. Let us assume that the working voltage is 2·17V, that is, a 15% shift from 2·55V. The minimum load working point now corresponds with a thermistor series resistance of about 450Ω, and at full load the input voltage will have fallen to about 1·75V.

Therefore, the change (δV_o) in bridge voltage from 1/5 load to full load is given by

$$\delta V_o = (2{\cdot}17-1{\cdot}75)\text{V} = 0{\cdot}42\text{V}.$$

The speed regulation from 1/5 load to full load is given by:

$$\text{Regulation} = \frac{0{\cdot}42}{2{\cdot}17}\times 100\% = 19{\cdot}5\%.$$

The measured regulation from 1/4 load to full load on an actual control system of this type was 18%. The agreement is therefore quite good.

If a system were being designed it would not be possible to arrive at the working voltage sufficiently reliably to estimate regulation in the above manner. It is, however, possible to arrive at a circuit design in another way. From the control characteristics in Figs. 16 and 17, further graphs can be plotted of control sensitivity in degrees per volt at the input terminals as a function of firing angle (Figs. 18 and 19). Inspection of Fig. 19 shows that, with this circuit arrangement, 2V represents the maximum usable working voltage at large firing angles. Above this voltage, at large firing angles, the sensitivity increases rapidly, and so the operation of the system may be erratic or unstable. It is therefore desirable to design the circuit aiming at a bridge output voltage of around 2V for the low-speed end of the range. This will result in a circuit similar to that of Fig. 15. Having selected this working region it can be seen that the input sensitivity at large firing angles (low speed) will be about 50 degrees per volt. If the firing angle range to be covered from minimum to maximum load is, say, 20°, then the control voltage change (δV_o) is given by,

$$\delta V_o = \frac{20}{50} = 0{\cdot}4\text{V}.$$

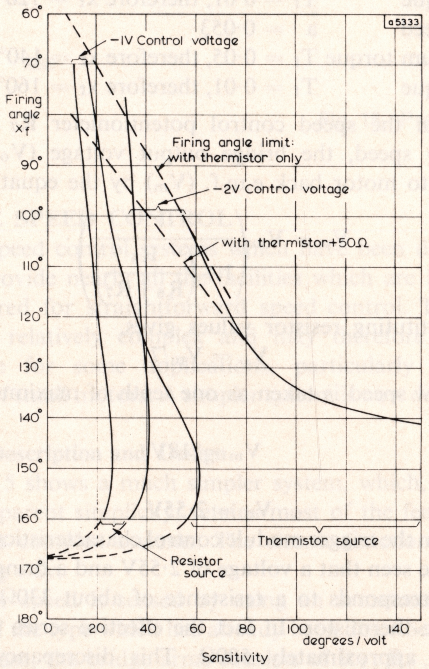

Fig. 18—Control characteristics for MY5001 using terminals 9 and 10 with shorting link between terminals 11 and 12

The motor speed regulation over this load range will be given by,

$$\text{Regulation} = \frac{0{\cdot}4}{2} \times 100$$
$$= 20\%.$$

It is interesting to note that this second approach gives nearly the same result as the first, but requires far less accurate initial data. The circuit constants have now to be selected to produce the correct working region.

The same procedure can be adopted to find the regulation at maximum speed. In this case however a larger working voltage is permissible. At 2·75V the control sensitivity is about 65 degrees per volt and the firing angle range about 25°. Therefore

$$\delta V_o = \frac{25}{65} = 0{\cdot}385\text{V},$$

and

$$\text{Regulation} = \frac{0{\cdot}385}{2{\cdot}75} \times 100$$
$$= 14\%.$$

The circuit in Fig. 15 gave a regulation of 10% at full speed, a rather better performance than would be achieved at the above working point. A small improvement would be achieved by shifting the working point to a slightly higher voltage, but care must be taken not to move into a potentially unstable region, a point which will be referred to again a little later.

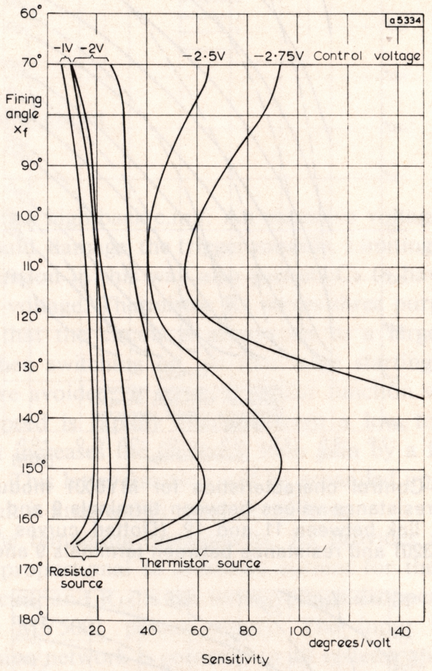

Fig. 19—Control characteristics for MY5001 using terminals 9 and 10 with 180Ω between terminals 11 and 12

To complete the design from the point reached so far, we have to arrange that the armature voltage bridge outputs (V_o) at low and high speeds are as follows.

At low speed; $V_o = 2{\cdot}0\text{V}$.
At high speed; $V_o = 2{\cdot}75\text{V}$.

But we also know that the motor back e.m.f. at high speed is 180V. Therefore at high speed the ratio of the bridge arms should be given by the following equation.

$$\frac{R_6}{R_6+(R_7+R_8)} = \frac{2{\cdot}75}{180} = \frac{1}{65}.$$

At low speed, the equation is as follows.

$$\frac{(R_6+R_7)}{(R_6+R_7)+R_8} = \frac{2{\cdot}0}{k\,.\,180},$$

where

$$k = \frac{\text{low speed}}{\text{high speed}}.$$

If $k = 0{\cdot}1$, then, at low speed,

$$\frac{(R_6+R_7)}{(R_6+R_7)+R_8} = \frac{1}{9}.$$

If (R_6+R_7) is, say, 550Ω (with $R_6 = 50\Omega$ and $R_7 = 500\Omega$), then $R_8 = 4400\Omega$,
and, at high speed,
$R_6 = 69\Omega$.

In a practical design the nearest preferred values would be used, giving,

$R_6 = 68\Omega$
$R_7 = 500\Omega$ (potentiometer)
$R_8 = 3{\cdot}9\text{k}\Omega$.

These values are near those given for the practical circuit in Fig. 15 where,

$R_6 = 47\Omega$
$R_7 = 500\Omega$
$R_8 = 3{\cdot}3\text{k}\Omega$.

These component values permit a maximum speed a little in excess of the rated speed at no-load, in order to give the full rated speed at full-load torque; hence R_6 has the lower value. At the low-speed end of the range the minimum speed has been set at rather less than 1/10 maximum speed by selecting the next lower preferred resistor for R_8. This design procedure may be used to design speed control systems with a wide range of characteristics; the above example serves to illustrate the method and the degree of correlation with a practical example. The measured performance of an actual circuit is shown in Fig. 20.

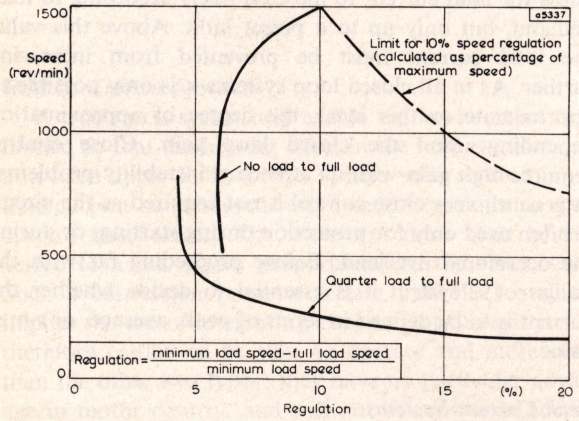

Fig. 20—Regulation versus speed for simple speed control circuit

When designing a circuit of this type, certain general principles should be used as a guide. Resistor R_4 should not be too large, since it has to carry the full armature current; a value between 0·1 and 0·2 times the armature resistance is a good compromise between good control and a not-too-high power dissipation. Resistor R_7 should not be more than about 500Ω; this then determines the value of R_8. It may be an advantage to make R_7 smaller, but R_8 has then to have a smaller value with a corresponding increase in dissipation. The resistance values suggested in Fig. 15 represent a reasonable compromise. Having fixed the value of R_4, and hence the voltage drop across it, the diode string D_5, D_6, D_7 is chosen so that it starts to conduct at the full-load armature current. If R_4 is about 1/7 of the armature resistance, three diodes will be required; if R_4 is 1/10 of the armature resistance only two diodes will be necessary (see components list).

The capacitor C_2, together with resistors R_6 and R_7, constitute the a.c. balance components of the bridge as described earlier. The time constant $C_2(R_6+R_7)$ must be made the same as that of the motor armature, thus:

$$C_2 = \frac{L_a\,.\,10^6}{r_a(R_6+R_7)}\,\mu\text{F},$$

where L_a is the armature inductance and r_a is the armature resistance.

Reference was made earlier to a possible risk of instability if the working point is chosen too high; this may arise for the following reason. The MY5001 trigger module is based upon a very simple RC timing circuit which is also widely used in similar circuits using pnp/npn transistor pairs and unijunction transistors. With this circuit provision has to be made to discharge the timing capacitor at the end of each supply half cycle, if the delay time for the following half cycle is to be correct. This is normally done by triggering the circuit at the end of the cycle so that the capacitor is discharged into the load, which, in this case, is the thyristor gate cathode circuit.

At certain delay values, where the pulse repetition rate is nearly an integral number per cycle, the capacitor voltage will very nearly reach its maximum value just at the end of the half cycle. The whole energy will therefore have to be dissipated in the thyristor gate circuits. It has been found in practice that if the thyristor gates are both of high impedance, it may not be possible to completely discharge the capacitor before it starts to recharge for the next half cycle.

If this is so, then the next firing pulse will appear early. A small change in delay about this critical value may result in a step change in delay for the next half cycle. The magnitude of these jumps in delay is minimised if the thyristor gate impedance is low, but this still has a noticeable effect when they occur in the region where the rate of charge of the capacitor is low, that is, to the left of the characteristics of Figs. 16 and 17. It is for this reason that a working point too far to the left of these characteristics should be avoided. Fortunately the effect of these discontinuities, when they do occur, is not too serious in this particular application as a phase advance on one half cycle is normally accompanied by a corresponding phase retard on the next, so that the mean power supplied to the motor is not affected. The only indication is that a small 50Hz ripple component appears superimposed upon the normal 100Hz ripple under certain operating conditions. This will normally be observed most markedly at low speed, with a moderately light load. The effect also shows up while the motor is accelerating under current limit control.

Current Limit

The current limit module type MY5051 is described in detail in the module data, and so a full description will not be repeated here. The output transistor of the module is connected in series with the charging resistance chain of the trigger module RC timing circuit. Normally this transistor is bottomed and therefore has a negligible effect upon the circuit operation, but when the input to the module exceeds a certain critical voltage, about 3V, the transistor begins to turn off and, in so doing, delays the firing angle of the trigger module. In a simple circuit of this type where cost is a first consideration the motor current will normally be sensed with a low value series resistor shown as R_3 in Fig. 15. The voltage across this resistor will be proportional to armature current, which consists of unidirectional current pulses. The exact waveform will depend upon motor speed and load.

To ensure that the current limit protection operates fast, to give complete protection under all normal operating conditions, the charging time constant of the integrating capacitor in the current limit module has been made about 10ms. However, to obtain good integration, and to stabilise the system, it is necessary to make the discharge time constant at least 100ms. Under these circumstances the input circuit operates as a peak sensitive system and responds to the peak repetitive armature current. The motor torque is proportional to average current and as the armature waveform changes under different speed and torque conditions, so the relation between peak and average current change. As the motor is loaded and the speed falls, the flywheel diode conduction period will increase to that the current waveform will become less peaky and the peak-to-mean ratio will approach unity. The effect upon the current limit operation will be that the limiting action will operate at progressively higher torques as the speed falls and the circuit will behave as though the control loop gain is very low. This appears to be an unavoidable consequence of using a fast-operating, simple circuit of this type.

To ensure that the current limit operates at a definite torque the circuit must be made to respond to the average value of armature current. In view of the discontinuous current waveform, particularly with a single phase system, a fairly long time constant integrating circuit must be used, and this inevitably makes the current limit circuit slow to respond to sudden changes in load current.

The current value at which current limit starts to operate is determined by the value of resistor R_3. The current limit module starts to turn off when the peak voltage across R_3 reaches 3V. The correct value for R_3 can be estimated as follows. Current limit will normally be required to operate at just over full-load torque. At this point the armature current waveform will be approaching a full-wave rectified sine-wave but will be discontinuous to the extent that the peak-to-mean ratio will typically be about 2, therefore for a peak voltage of 3V the mean value will be about 1·5V. Allowing for current limit to start a little over full-load, the value of R_3 may be estimated approximately from the following empirical equation.

$$R_3 = \frac{1}{I_{max}}\Omega,$$

where I_{max} is the rated full-load armature current in amperes.

In certain circumstances it may be found that the value obtained in this way is either a little high or a little low, as the waveform will differ to some extent for different motors.

In these cases it is a simple matter to slightly alter the value of R_3 when testing the prototype. Once the correct value has been found by trial it can be used for all other motors of the same type.

PERIPHERAL EQUIPMENT

With the electronic control circuits already described it is relatively simple to add other protection circuits. These are described separately below.

Current Limit

Reference has already been made to a number of aspects of current limit. The various circuit diagrams show how this feature has been included in the three systems which have been described. Reference has also been made to some effects of waveform upon the operation of a current limit circuit. The purpose of this section is to outline some other general features of these and other systems and to give some detailed advice about circuit design.

In principle the purpose of a current limit system is to allow the load current to increase freely according to load demand, but only up to a preset limit. Above this value the load current must be prevented from increasing further. As in all closed loop systems it is only possible to approximate to this ideal, the degree of approximation depending upon the closed loop gain. Close control requires high gain with its attendant instability problems; in general, very close control is not required as the circuit is often used only for protection during starting, or during the occasional overload. Before proceeding far with the design of a system it is essential to decide whether the current is to be defined in terms of peak, average, or r.m.s. value.

Peak Current Systems

Peak current systems are fast to operate; they are able, at their best, to operate cycle by cycle. They therefore provide the fastest possible protection, but unless precautions are taken they are liable to suffer from instability. A true fast system will have little or no 'memory', and is therefore liable to alternate between 'on' and 'off' states. It often does this at supply frequency, and so the thyristor bridge 'half-waves', that is, it conducts on one half-cycle and blocks on the next. This effect can be overcome and the stability of the system improved by making the current limit circuit fast to operate and slow to recover, by having a short charge time constant and a long discharge time constant. This technique has been used in the MY5051 current limit module. The input to a peak system can be obtained either from a current transformer in the a.c. line, or from the d.c. output by using a series resistor or d.c. current transformer. In all cases the system performance will be the same since the peak current in both parts of the circuit is identical; the thyristors always turn off after the peak current point.

When a peak current system is used with a d.c. motor control system the motor torque in the current limit region will follow a law determined by the peak-to-mean ratio of the armature current waveform. At stall, the peak-to-mean ratio will approach unity, and may be about 1·1. At high speed and full load it may be between 2 and 3. Thus if current limit is set to just operate at high speed and full load, the torque may increase to between 2 and 3 times this value before the motor stalls.

Mean Current Systems

If the current limit circuit is to respond to the mean load current some kind of integrating circuit must be employed. The integration must be performed over a number of cycles if it is to be effective; therefore such systems must be slower to operate than peak systems. In most cases a mean current system will be required to respond to the mean load current; therefore current sensing must be by series resistance or d.c. current transformer in the load circuit. In the case of a true mean current system, the motor torque will remain constant in the current limit region. The MY5051 module can be used as a mean current limit circuit by either adding resistance in series with the input terminal or providing a separate integrating circuit external to the module.

R.M.S. Current Systems

A true r.m.s. system will enable the load current to be limited so that the heating effect in the load is maintained at a fixed upper limit. Such a system will require that the load current be measured with a small resistance or a d.c. current transformer. The waveform must then be converted to a d.c. signal proportional to the r.m.s. value. Such a conversion presents certain difficulties and also implies the use of integrating circuits. R.M.S. systems are therefore certain to be slow to operate and more costly than the other two types; they have no particular advantage in motor control, and will not be considered further.

Application to Motor Control

Only peak and mean current systems will be considered for the reasons just given. For small motors, up to say 1hp, the peak system is more suitable; it is fast and provides good protection. Small motors are often operated near their upper torque limit and may be required to deliver peak starting torques of 2 or 3 times full rated torque. The peak system permits this condition, but for full system protection the thyristor bridge should be rated at about twice motor full-load current, and a slow operating current trip or thermal overload protection incorporated to protect the motor in case it should remain in the stalled overload condition for too long.

For larger motors it may be uneconomic to provide a thyristor bridge with a large overload capacity, particularly if the motor is never required to deliver large starting torques. Under these conditions a mean level system should be used and precautions taken to avoid rapid changes in load which might exceed the current limit response rate. A suitable 'slow start' circuit is essential under these conditions. A thermal overload trip may still be necessary because the mean-to-r.m.s. current ratio may be at least 1·2 under full-load conditions, and so limiting the current to a maximum average value may not provide full protection for the motor.

If the fast response of the peak system and the control characteristics of the mean system are both required the two circuits can be combined by feeding control signals from both into the adding junction at the appropriate voltage and gain levels.

Current Sensing

The MY5051 current limit module requires an input of about 3V before it starts to operate, and it becomes fully operated with an input of about 4V.

If the load current is not too large, this input voltage is best provided by a resistance (R) in series with the load. The practical limit is reached when the power dissipation of this resistor becomes excessive.

For a peak current system the resistance in ohms is approximately given by the equation,

$$R \simeq \frac{1}{\text{Mean Full Load Armature Current in amperes}}\Omega.$$

For a mean current system the resistance will depend upon the attenuation of the integrating circuit. In most cases this will be about 2, and so the resistance will be given by the equation,

$$R \simeq \frac{6}{\text{Mean Full Load Armature Current in amperes}}\Omega.$$

A current transformer in the a.c. line is really suitable only for use with a peak current system, in which case the peak output voltage should be 3V at full-load current. For convenience in specifying the transformer this will correspond with about 2V r.m.s. output at full-load r.m.s. sine-wave input.

If a d.c. current transformer (which gives an output waveform very nearly the same as that which would be obtained by using a series resistor) is used, the design procedure is the same as for a resistor. This type of transformer is suitable for either peak or mean current systems. It does not suffer from the high power dissipation of a resistor, and at the same time provides isolation while retaining the d.c. level of the primary current in the secondary circuit. The current transformer will however be more expensive than a resistor with smaller currents.

The circuits shown in Figs. 7 and 11 both use an a.c. current transformer with a compromise between peak and mean systems to give reasonably fast response. As an alternative, either a resistor or a d.c. current transformer may be connected in series with the positive terminal of the thyristor bridge and the top of R_2. A peak or mean control voltage may then be taken from this component.

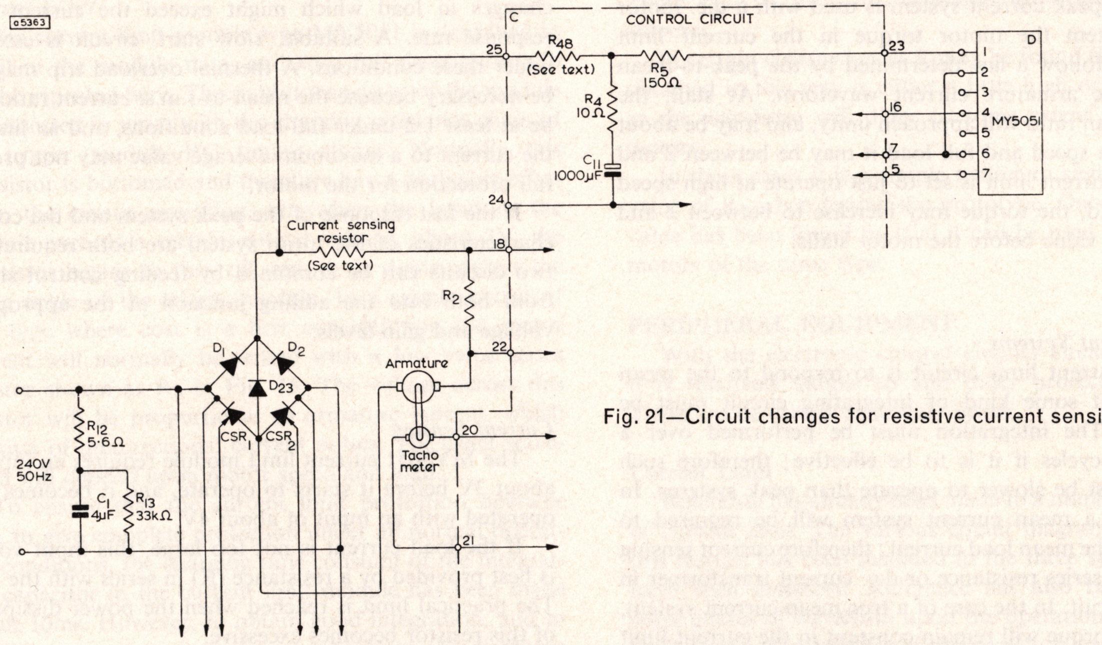

Fig. 21—Circuit changes for resistive current sensing (see text)

This and other minor circuit changes are shown in Fig. 21. The value of R_{48} should be 220Ω for a mean level circuit, or it can be reduced to zero when used with an a.c. current transformer to give a faster operating system.

Starting Circuits

When a d.c. shunt motor is stationary, the armature impedance may be very low and is often less than 1/20 of normal full-load value. It is therefore essential that the starting current is limited to a safe value while the motor is accelerating. After the first few half-cycles of the supply a fast current limit circuit can provide the necessary limiting action, but initially some other means must be used.

In the simple circuit in Fig. 15, the thermistor provides this protection. As it heats up, it allows the trigger module firing angle phase to advance and the current limit circuit to take over control. It may be important in some applications to ensure that the starting circuit has a fast reset time so that it can respond to repeated starts; the thermistor circuit is particularly good in this respect. Three conditions can arise when starting; these are,

(i) Normal start with normal load; motor accelerates fast to running speed.

(ii) Start with heavy load, or with locked motor because of a mechanical fault.

(iii) Rapid restart in condition (i) before speed has built up, or in condition (ii) while motor is still locked.

Condition (i) is normal and presents no problems; if the motor is turned off and on rapidly, inertia will maintain speed sufficiently to prevent large surges. If the motor is turned off and then stopped, sufficient time will have elapsed for reset, which for the thermistor is 100ms.

Condition (ii) is similar to (i) for the first start as there is no instantaneous difference between a locked motor and a normal inertia load. After the first few half-cycles the current limit circuit would become fully operative.

In condition (iii), where a repeat start follows closely after a full current limit condition, the system is safe provided that the current limit recovery (or capacitor discharge) time is much greater than the thermistor cooling time. Tests show that if the circuit is switched off and on with a delay value mid-way between the current limit discharge time constant and thermistor time constant, the current overshoot is only about 50% above the normal current limit value for 2 or 3 half-cycles; this represents the worst possible conditions, and is acceptable.

The starting circuits used in the two more complex circuits (Figs. 7 and 11) make use of an RC network to provide an initial phase delay for the firing pulses. No special steps have been taken to rapidly reset these circuits, and so, although they provide complete protection under starting conditions (i) and (ii), a certain time must elapse before restarting under condition (iii) if complete protection is required; this time may be as long as 5s.

The slow start circuit operation depends upon the slow discharge of C_{12} when the supply is remade; it is therefore essential that the mains supply to the control circuit transformers should be broken whenever the supply to the thyristor bridge is disconnected. It is recommended therefore that wiring diagrams should be drawn with the control transformer primaries linked directly to the thyristor bridge supply. In this way there will be less risk of separate switching becoming incorporated in this part of the circuit.

Open Circuit Field Protection

If the separately excited field of a d.c. shunt motor fails to energize, no back e.m.f. will be developed by the armature and very large currents will flow. In the case of a thyristor controlled armature supply having a current limit circuit, the current will be limited to a little over the full load current. Under these conditions the motor will slowly overheat, but little serious damage is likely to result, particularly if thermal overload protection is incorporated. In some cases however it may be considered desirable to arrange that the armature supply is cut off in the absence of the correct field current.

A simple addition to the field current compensation circuit which is described earlier, on page 89, provides the additional facility of open circuit field protection. If the field current falls below some predetermined value, say 70% of normal, there will be insufficient voltage at the tapping point of R_{15} (Fig. 11) to enable the voltage regulator diode D_{15} to conduct. Transistor TR_3 will therefore cease to conduct, and the full supply voltage will appear at its collector. This voltage is sufficient to make the voltage regulator diode D_{20} conduct and by-pass the high resistance R_{28}. The remainder of the output voltage from TR_3 is then applied, with very little attenuation, through the low resistance R_{30} to the amplifier input, driving it into its 'end stop', or 'off', condition.

Mains Transient Suppression

It is well known that there may be voltage transients superimposed upon the a.c. mains supply. The magnitude and energy level of these transients, and their frequency of occurrence, cannot be predicted with accuracy, although a considerable amount of experience is accumulating with the increasing use of semiconductor rectifier and control equipment. It is considered to be good practice to use rectifiers and thyristors with peak voltage ratings of twice the peak supply voltage, that is, 700V for a 240V r.m.s. single phase supply and 1200V for a 415V three phase supply. In some cases where initial cost is a first consideration these voltages may be reduced to some extent; in others, where reliability is vital, higher voltages may be considered to be desirable. Whatever the factor of safety chosen, it is not possible, for cost reasons, to use devices which will block all possible transient voltages. Avalanche devices help considerably to make transient suppression unnecessary, particularly in simple rectifier equipment, but thyristor control equipment presents different problems. For example, if a transient should cause a thyristor to conduct in the forward breakover condition when the control system is intended to be off, an excessive current may flow so that a thyristor may stay on for the remainder of the half-cycle. The magnitude of this current may be sufficient to cause damage, or the impulse may cause a piece of machinery to rotate. It is therefore good practice to use adequate transient suppression in all cases where controlled rectification is used.

In principle, a capacitance across the supply terminal will absorb the high voltage lower energy transients; there are, however, circuit details which must be considered before a suitable circuit can be designed.

The supply cable will in most cases appear at the output terminals as a transmission line which, at frequencies below about 50kHz, looks largely inductive. A capacitance placed directly across the terminals may 'ring' with this inductance, and produce a damped oscillation (as a result of switching) with an amplitude considerably in excess of that which would be present in the absence of the capacitance; voltages of the order of 2kV have been measured on a 240V single phase system. It is therefore essential to incorporate series resistance with the capacitance to at least critically damp the circuit to prevent ringing.

In addition to the suppressing of transients, a suitable capacitor serves another essential function. When a thyristor controlled bridge is connected to the mains supply by a mechanical contactor the instantaneous supply voltage will be applied to the forward connected thyristor with a very high rate of rise. If contact is made at any point in the cycle, other than quite near the zero cross-over points, the value of dV/dt at the thyristor is likely to exceed the maximum permitted value, so that the thyristor may turn on. This undesirable effect can be avoided by a suitable choice of input shunt capacitor, which together with the series line impedance, will limit the maximum rate of rise to a safe value. It is important in this case that the capacitor be provided with a reasonably short time-constant discharge path, so that it is ready for re-use should switching operations follow each other in quick succession.

Measurement and calculation indicate that the average low-power single phase supply (up to 30A) has a series inductance of about 50μH. With such a supply a shunt capacitor of 2μF and a series resistor of 5Ω results in a maximum dV/dt of 20V/μs. A 4μF capacitor with 5Ω reduces dV/dt to 13V/μs. Thyristors will not normally turn on with these low values of dV/dt. In cases where it is thought that the mains impedance may be lower than 50μH, it is suggested that an a.c. choke of between 50 and 100μH be connected in series with the supply line. It should be remembered however that, in this case, the supply will no longer look like a transmission line so the value of dV/dt may be different. In higher power and three phase circuits it is considered to be good practice to fit small series chokes in general purpose equipment.

Motor Thermal Overload Protection

Positive temperature coefficient (p.t.c.) thermistors are available with very high values of temperature coefficient above certain specified temperatures. These temperatures have been selected to lie within the range of upper temperature limits for motor windings. If a suitable p.t.c. thermistor is embedded in the windings of a motor it provides a convenient and direct method of indicating when the windings reach the maximum safe temperature limit. If the p.t.c. thermistor is included in a suitable part of the control circuit, the rapid change in resistance at its critical temperature can be made either to turn off, or to phase back, the trigger circuit; this may be temporarily until the motor cools down, or permanently until manually reset. The circuit modifications are shown in Figs. 22 and 23.

Tachometer Speed Control with P.T.C. Protection

The circuit description which follows applies to the circuit in Fig. 7 and also the armature voltage control

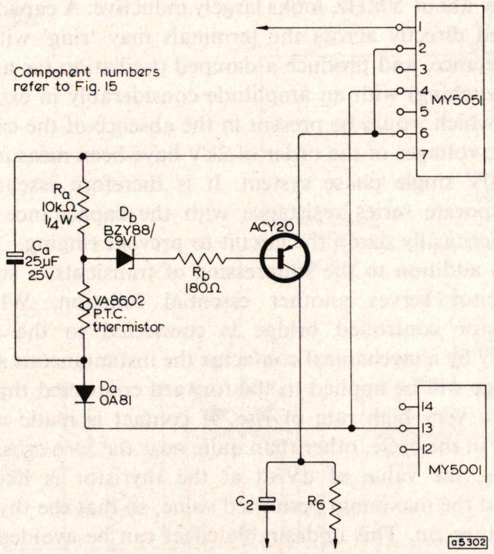

Fig. 22—Additional components, for the circuit of Fig. 15, to provide p.t.c. thermistor protection

circuit in Fig. 11 which, in this respect, are identical. The circuit details are shown in Fig. 23.

At normal operating temperatures the resistance of the p.t.c. thermistor is low, and the voltage reference diode D_b does not conduct. When the critical temperature of the p.t.c. thermistor is reached, its resistance rises rapidly, the diode D_b conducts, and the amplifier is driven to the off state. The capacitor C_a and diode D_a are included for the following reason. When the p.t.c. thermistor cools, the diode D_b ceases to conduct and the amplifier turns on again with only a very small change in temperature of the p.t.c. thermistor. Unless the p.t.c. thermistor is well bonded to a mass of large thermal capacity, the rate of change of resistance will be fast enough to exceed the response time of the current limit circuit, and since the slow start circuit will have been by-passed, a large current surge is possible. The additional diode D_a and capacitor C_a constitute a secondary slow start circuit to cover this contingency. When a typical p.t.c. thermistor type VA8602 is used with this circuit the speed control circuit functions normally below 118°C and is completely off above 119·5°C.

Simple Speed Control with p.t.c. Protection

The simplest possible arrangement which can be used with this circuit requires no additional components. The p.t.c. thermistor should be connected in series with the thermistor R_9 (Fig. 15) between terminal 15 of the control circuit and terminal 9 of the MY5001 trigger module. Although this simple arrangement provides entirely satisfactory protection, it does limit to some extent the maximum voltage which can be obtained from the thyristor bridge. An alternative arrangement which does not suffer from this disadvantage is shown in Fig. 22. This circuit also gives a rather closer control of temperature. Using p.t.c. thermistor type VA8602 the circuit functions normally below 120°C and is completely off above 125°C. The circuit operates in a very similar way to that of Fig. 23 except that the capacitor C_a and the diode D_a serve a completely different function. The low voltage supply in the simple circuit consists of a sinewave clipped by the voltage regulator diode. Negative going pulses at the end of each half-cycle are used for synchronisation and have therefore to be preserved. The diode D_a and capacitor C_a provide a smooth d.c. supply for the p.t.c. thermistor circuit and at the same time isolate this voltage from the trigger module supply. A separate slow start circuit is not required in this case as the thermistor R_9 in the motor control circuit provides this in all circumstances.

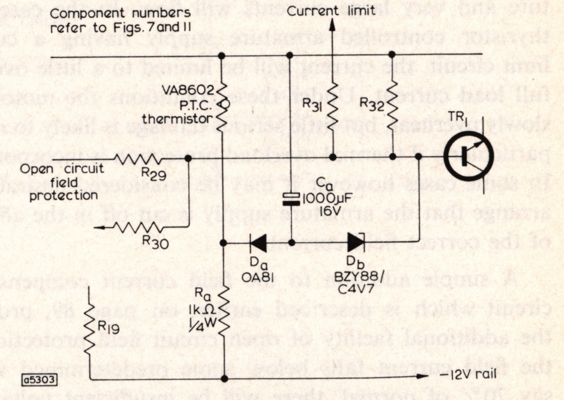

Fig. 23—Additional components, for the circuits of Figs. 7 and 15, to provide p.t.c. thermistor protection

PRACTICAL CIRCUIT DESIGN AND COMMISSIONING

The notes given in this section are intended as a guide to those concerned with the practical application of the circuits described in this article.

Before discussing each circuit individually the selection of a suitable motor for electronic speed control warrants some consideration. This subject is dealt with in considerable detail in Ref. 2, but, in most cases, there is no need to delve into this in great detail. If the motor is purchased from one of the manufacturers who specialise in motors of this type, then most of the details will have been taken into account by them. It is then necessary only to specify that the motor is to be speed-controlled with a single-phase thyristor bridge, and give the following data:

(i) Horse power at rated speed.
(ii) Maximum speed.
(iii) Armature voltage, usually about 170 volts d.c.
(iv) Field voltage, usually about 200 volts d.c.

In addition the motor makers should be able to specify the following data:

(v) Armature average current at full load.
(vi) Field current.
(vii) Armature resistance.
(viii) Armature electrical time constant.

Simple Circuit (Fig. 15)

Assuming that a suitable motor has been selected, first find the motor armature resistance, either from the manufacturer, or by measurement. A voltmeter, ammeter and low voltage d.c. supply may be used if a suitable bridge is not available. For example if a 1hp motor has a resistance of 1Ω, then a 6V accumulator across it will give 6A. Any reliable moving coil meter will give a sufficiently accurate measurement.

Having obtained the motor resistance, calculate the value of R_4 (Fig. 15) by the equation,

$$R_4 = (\text{Armature Resistance}/7)\Omega.$$

From the motor rating-plate, find the full-load armature current.

The approximate value of R_3 (Fig. 15) can be obtained from the equation,

$$R_3 = (1/\text{Full-load Armature Current in Amps})\Omega.$$

The value of R_3 obtained in this way is only approximate; the exact value is best found by measurement on the complete system.

The resistors R_3 and R_4 will be of low value, and consequently, may not be available as standard components; they can easily be constructed from a length of heavy gauge resistance wire using the manufacturer's data to obtain the correct length of wire for the required resistance. This procedure will usually enable a suitable component to be made with sufficient accuracy; the wire data will also give the recommended maximum current for each wire gauge.

After determination of the values R_3 and R_4, and taking a typical value for C_2 as 25μF, the circuit may be wired up. First set the speed control to the low speed setting, that is, with the slider of R_7 at the end towards R_8, and then connect the mains supply. The motor should now rotate on no load at about 1/6 full speed and the speed should be controllable up to about 20% above nominal top speed. If the direction of rotation is wrong, reverse the field connections.

The motor should now be tested on load with an ammeter in the armature circuit so that a check can be made on armature current.

The motor speed regulation against load torque can be measured, and should be better than 10% from no load to full load over a 10:1 speed range, except at low speed, where the regulation at very low torques is not so good. At low speed the measurement is therefore best made from, say, 1/4 full load torque to full torque.

If the motor does not behave satisfactorily it will probably be for one of the following reasons:

(i) Resistor R_3 may have too high a value. This will result in a premature fall in speed as torque is increased resulting from early onset of current limit; this is particularly likely at higher speeds. Check by short circuiting part of the resistor until the correct performance is achieved.

(ii) Capacitor C_2 may have the wrong value. If it is too large the system may oscillate, that is, the motor will hunt, particularly at low speed. The remedy is to replace the capacitor with the next lower preferred value.

If C_2 is too small, the motor regulation will be poor both at low and at high speed; it is inadvisable at this stage, however, to increase the value of C_2 without checking the bridge output with an oscilloscope. Connect an unearthed oscilloscope across C_3; the ripple voltage should be less than about 0·25V. If it is greater than this try larger or smaller values of C_2 to see if the ripple can be decreased in amplitude. Select the capacitor which gives the minimum ripple with a lagging phase shift from the supply.

(iii) The resistor R_{12} may not have the correct value. Increasing R_{12} will improve regulation but if it has too high a value the circuit may be unstable, especially at low speed and low load. The maximum value of R_{12} should not exceed 200Ω.

Armature Voltage Speed Control (Fig. 11)

Motor data can be obtained as described for the simple system, and the value of R_2 calculated from the equation,

$$R_2 = (\text{Armature Resistance}/10)\Omega.$$

The field compensation component values for a 240V supply can be calculated from the following equation,

$$R_{14} = (\text{Field Resistance}/64)\Omega,$$

and,

$$R_{15} = (\text{Field Resistance}/32)\Omega.$$

If current limit is obtained from a series resistor (R_x) in the armature circuit then,

$$R_x = (5/\text{Full Load Armature Current in Amperes})\Omega,$$

and

$$R_{48} = 220\Omega.$$

If current limit is obtained from a current transformer, see 'Current Sensing' on page 95. In this case R_{48} may be reduced to zero if faster response is required.

The five preset resistors enable the following circuit functions to be set up; they are,

R_{10}, Maximum speed,
R_{15}, Field current compensation,
R_{22}, Minimum speed,
R_{44}, End stop,
R_{47}, IR compensation.

These resistors can all be set using only the motor and a separate tachometer as follows:

(i) Set R_{15} with its slider at the end nearer to terminal 7.
(ii) Set R_{22} to zero resistance.
(iii) Set R_{47} to maximum resistance.
(iv) Set R_{23} (speed control) to the low-speed setting, that is, with the slider at the end of the track towards R_{22}.
(v) Turn on the supply and adjust R_{44} until the motor just runs smoothly on no load. Avoid too low a speed setting (where firing pulses may be erratic) by listening for a clear 100Hz hum from the motor.
(vi) Adjust R_{22} to give the correct low speed. If the speed range is large it may be necessary to mechanically load the motor to say 1/5 of full load torque before doing this.

The reason for this is as follows. The trigger module cannot produce reliable trigger pulses at a firing angle greater than about 170°, because the amplifier end-stop (diode clamp D_{21}, D_{22}) is set to exclude operation in the region between 170° and 180°. A motor may be able to rotate on no load at a speed of at least 1/25 of full speed at this firing angle, although quite a small additional load will drop this to perhaps 1/100 full speed, after which the normal feedback speed stabilising circuit will take control.

(vii) Set the speed control R_{23} to maximum speed.

(viii) Adjust R_{10} to give the correct full speed.

(ix) Set the speed control R_{23} to low speed again.

(x) Set the IR compensation gain with R_{47} so that the fall in speed from about 1/5 full torque to full-load torque is between 2 and 3 rev/min at low speed (about 15 to 20 rev/min for a 1500 rev/min motor or pro rata for other top speeds). An armature current ammeter may be used as an indication of torque if torque measuring equipment is not available. The motor rated full-load current may be taken to correspond with full-load torque. Armature current will than give a nearly linear indication of torque below this level.

For smaller sizes of motor (below 5hp) a simple, effective brake can be quickly and cheaply constructed by fixing a pulley on the motor shaft, clamping the motor to a bench, and using two lengths of timber as a pincer on the pulley.

(xi) Turn off the supply and temporarily connect a resistor in series with the field; this resistor should equal about half the field resistance, and will drop the field current to about 67% of normal value.

(xii) Turn on the supply and with the motor running at about two thirds of full speed adjust R_{15} until the motor just begins to slow-down. It is essential that the preset control should not be turned back to restore the motor to normal operation without first turning off the mains supply; slow start does not operate in this condition.

When R_{15} is set as recommended above, the motor speed will be approximately compensated, for field current changes of ±10% at two thirds of full speed; armature current will be reduced to a very small value if the field current falls below 67% of its nominal value.

To obtain optimum performance it may be necessary to slightly modify the value of R_{28}. It is convenient to make part of this resistance variable until the best value has been determined.

(xiii) Should the motor show signs of instability, or hunting, particularly at low-speeds, check that the value of C_2 is correct using the technique described for the simple system on page 99. In this case, connect an unearthed oscilloscope between the junction of R_3, D_8 and the junction of C_2, R_{21}.

Tachometer Speed Control (Fig. 7)

The setting up procedure for this circuit is very similar to that described above for the armature voltage control circuit. In this case, however, there is no armature series resistance to design, and the field feedback circuit is only used for open-circuit field protection. Consequently, R_{28} is omitted and does not therefore need adjustment. Again, for a 240V supply

$$R_{14} = (\text{Field Resistance}/64)\Omega,$$

and,

$$R_{15} = (\text{Field Resistance}/32)\Omega.$$

The tachometer reference voltage is nominally 14V; a d.c. tachometer should therefore be selected which has an output voltage of approximately 14V at maximum motor speed. If a higher voltage tachometer is used the output can be reduced with a potential divider or additional series resistance, provided that the effective source resistance of the generator does not exceed 1kΩ.

Normally, a tachometer control system will only be selected if high speed setting accuracy and long term speed stability are required; otherwise the armature voltage control system, with additional IR compensation if necessary, will give an equally good regulation performance. If a tachometer is used primarily to obtain high accuracy it is therefore important that the tachometer itself should have the required stability. Many tachometers will have the required stability when run at high speed; for example, a 10V per 1000 rev/min tachometer only needs to have an output stability of ±100mV if it is to be accurate to ±1% at 1000 rev/min. If, however, the speed is dropped to say 1/20 top speed (50 rev/min), then a ±1% stability demands a maximum output drift of ±5mV. The variation in commutator-to-brush voltage drop, alone, is very likely to be greater than this unless special precautions are taken. It will therefore be evident that care must be taken, when selecting a tachometer, to ensure that the electrical properties are suitable. If a very wide speed range is required it is better to use a tachometer with a high output voltage, say 100V per 1000 rev/min for a 1500 rev/min motor, so that, even at low speed, the tachometer output voltage is larger than the various contact and thermal potentials around the control loop.

In addition to the electrical properties of the tachometer, the mechanical properties should also be carefully considered. Some of the lower priced tachometers available are primarily intended for instrument applications. If such instruments (for this is how they should be described) are mounted on a large machine and driven directly from the main shaft the transmitted vibration can quickly destroy their delicate bearings.

The coupling between motor and tachometer should also receive due consideration. It must be sufficiently flexible to relieve the tachometer bearings from strain but at the same time it must be torsionally rigid and without backlash. Any backlash or resilience at this point is almost certain to introduce hunting into the motor feedback loop. Hunting of this type may often be almost indistinguishable from that due to electrical or mechanical phase shift within the feedback loop. In the event, therefore, of instability (especially at low speed) it is advisable first to check the rigidity of the tachometer coupling before assuming that the trouble should be cured electrically.

Integration with External Controls

In certain circumstances, it is necessary to change motor speed, or start and stop the motor with some external control signal. In the armature voltage and tachometer systems the best point in the circuit to introduce such controls is at terminals 6, 8 and 15. For preset speed the potentiometer can be directly replaced by a switched resistor chain. For 'inching', the low speed limit resistor R_{22} can be set to give the 'inching' speed and a switch connected between terminals 6 and 20 to short-circuit this resistor for the stop condition.

Although short-circuiting R_{22} will, in most cases, stop the motor, a small armature current will still flow because of the end stop clamp on the amplifier output. The motor may therefore still rotate on very light load. If a true off condition is required it is recommended that the mains supply be disconnected from the thyristor bridge.

Alternatively, speed can be controlled entirely by an external voltage signal which should be connected between terminals 20 and 8 in place of the control resistor R_{23}. The voltage should be variable between 0 and 14V with terminal 8 positive with respect to 20. The control circuit will be driven to its end-stop condition when the control voltage is less than about 0·25V and to maximum output at 14V. The input resistance is normally 5kΩ but falls to about 4kΩ when the motor is running under current limit conditions. By controlling at this point, full starting and current limit protection is retained so that the motor can be instantaneously switched from rest to full speed with complete safety using only low voltage, low current control circuits. The source resistance of this control voltage should be less than 1kΩ.

In the case of the simple speed control circuit, it is not possible to give direct guidance about external control as this will depend upon the application, but the speed control resistor R_7 can be replaced by a switched resistor chain to give preset speeds. Open-circuiting terminal 5 will stop the motor. An inching control could consist of a switch to open-circuit the connection to terminal 5 for 'stop' or to connect to terminal 4 for 'inch'. Voltage control with this system is not normally possible.

APPENDIX 1

Components for Tachometer Control System (Fig. 7)

A printed circuit board intended for use with this circuit is available from Cressal Manufacturing Co. Ltd., Cheston Road, Aston, Birmingham.

B. Mullard thyristor stack (D_1, D_2, D_{23}, CSR_1, CSR_2)	Current rating 1·5 × motor full load current.
D. Current-limiting module	Mullard MY5051.
E. Trigger module	Mullard MY5011.
Transistors	
TR_1, TR_2	Mullard BC107.
TR_3, TR_4	Mullard BCY32.
Diodes	
D_3, D_4, D_5, D_6, D_7, D_8, D_9, D_{10}, D_{11}, D_{12}, D_{21}, D_{22}	Mullard BYX10.
D_{13}, D_{14}	Mullard BYX22–800 (or to suit field current).
D_{15}, D_{16}, D_{17}, D_{18}	Mullard BZY88–C4V7.
D_{19}	Mullard OAZ273.
D_{20}	Mullard BZY88–C8V2.
D_{24}	Mullard BZY95–C20 (see note).

Capacitors	
C_1	4μF, 1000V, paper.
C_3	200μF, 10V, electrolytic Mullard C428AR/D200.
C_4	40μF, 25V, electrolytic Mullard C428AR/F40.
C_5	1000μF, 16V, electrolytic Mullard C437AR/E1000.
C_8, C_9, C_{10}	400μF, 40V, electrolytic Mullard C437AR/G400.
C_{11}, C_{12}	1000μF, 10V, electrolytic Mullard C437AR/D1000.
C_{13}	80μF, 4V, electrolytic Mullard C428AR/B80.
C_{14}	10μF, 16V, electrolytic Mullard C428AR/E10.
C_{15}	400μF, 25V, electrolytic Mullard C437AR/F4000.
Resistors	
R_1	7·5Ω, 1W.
R_2	(1/10) × armature resistance, Ω.
R_3	33Ω, $\frac{1}{4}$W.
R_4	10Ω, $\frac{1}{4}$W.
R_5	150Ω, $\frac{1}{4}$W.
R_6	2·2kΩ, $\frac{1}{4}$W.
R_7	22Ω, $\frac{1}{4}$W.

R_8	560Ω, ¼W, high stability.
R_9	1·2kΩ, ¼W, high stability.
R_{10}	1kΩ, wire-wound, pre-set (variable).
R_{11}	390Ω, 1W.
R_{12}	5·6Ω, 3W, wire-wound.
R_{13}	33kΩ, 3W, wire-wound.
R_{14}	(1/64) × field resistance, Ω.
R_{15}	(1/32) × field resistance, Ω, wire-wound pre-set potentiometer.
R_{16}	470Ω, ¼W.
R_{17}	1kΩ, ¼W.
R_{18}	56kΩ, ¼W.
R_{19}	1kΩ, ¼W.
R_{22}	2kΩ, wire-wound pre-set (variable).
R_{23}	5kΩ, precision speed control
R_{24}	1·2kΩ, ¼W, high stability.
R_{25}	330Ω, 1W. (When D_{24} and R_{39} are included, reduce to 150Ω, ¼W.)
R_{26}	2·7kΩ, ¼W, high stability.
R_{27}	1·8kΩ, ¼W, high stability.
R_{29}	1kΩ, ¼W high stability.
R_{30}	5·6kΩ, ¼W.
R_{31}	6·8kΩ, ¼W high stability.
R_{32}	3·9kΩ, ¼W high stability.
R_{33}	12kΩ, ¼W.
R_{34}	1kΩ, ¼W.
R_{35}	15Ω, ¼W.
R_{36}	120kΩ, ¼W, high stability.
R_{37}	6·8kΩ, ¼W, high stability.
R_{38}	6·8kΩ, ¼W, high stability.
R_{39}	68Ω, 1W, or zero Ω (see note).
R_{40}	560Ω, ¼W.
R_{42}	82Ω, ¼W.
R_{43}	220Ω, 1W.
R_{44}	100Ω, wire-wound pre-set.
R_{45}	1kΩ, ¼W.
R_{46}	10kΩ, ¼W.
R_{47}	50kΩ, carbon pre-set Mullard type E097AD.
R_{48}	220Ω, ¼W.
VDR_1	Asymmetrical v.d.r. Mullard E295ZZ/01.

(There are no components numbered C_2, C_6, C_7, R_{20}, R_{21}, R_{28}, T_{41})

Current Transformer
Output voltage 2V into 7·5Ω with full load primary current.

Note:
Without additional stabiliser diode D_{24}, and with $R_{39} = 0\Omega$, the set speed changes by about 2·5% for a 20% change in supply voltage. With the additional voltage regulator diode D_{24}, and resistor R_{39}, the set speed changes by only 0·25% for a 20% change in supply voltage. Both measurements are for maximum set speed.

APPENDIX 2

Components for Armature Voltage Control System (Fig. 11)

A printed circuit board intended for use with this circuit is available from Cressal Manufacturing Co. Ltd., Cheston Road, Aston, Birmingham.

B. Mullard Thyristor Stack (D_1, D_2, D_{23}, CSR_1, CSR_2)	Current rating 1·5 × motor full load current.
D. Current-limiting module	Mullard MY5051.
E. Trigger Module	Mullard MY5011.

Transistors

TR_1, TR_2	Mullard BC107.
TR_3, TR_4	Mullard BCY32.

Diodes

D_3, D_4, D_5, D_6, D_7, D_8, D_9, D_{10}, D_{11}, D_{12} D_{21}, D_{22}	Mullard BYX10.
D_{13}, D_{14}	Mullard BYX22–800 (or to suit field current).
D_{15}, D_{16}, D_{17}, D_{18}	Mullard BZY88–C4V7.
D_{19}	Mullard OAZ273.
D_{20}	Mullard BZY88–C8V2.
D_{24}	Mullard BZY95–C20 (see note).

Capacitors

C_1	4·0μF, 1000V, paper.
C_2	(10^6/470) × armature time constant. μF, 50V. (Typically 25μF).
C_3	200μF, 10V, electrolytic, Mullard C428AR/D200.
C_4	40μF, 25V, electrolytic, Mullard C428AR/F40.
C_5	1000μF, 16V, electrolytic, Mullard C437AR/E1000.
C_7	0·22μF, 160V, paper, Mullard C296AA/A.
C_8, C_9, C_{10}	400μF, 40V, electrolytic, Mullard C437AR/G400.
C_{11}, C_{12}	1000μF, 10V, electrolytic, Mullard C437AR/D1000.
C_{13}	80μF, 4V, electrolytic, Mullard C428AR/B80.

Resistors

R_1	7·5Ω, 1W.
R_2	(1/10) × armature resistance, Ω.
R_3	33Ω, ¼W.
R_4	10Ω, ¼W.
R_5	150Ω, ¼W.
R_6	1kΩ, ¼W.

R_7	22Ω, ¼W.
R_{10}	1kΩ, wire-wound pre-set (variable).
R_{11}	390Ω, 1W.
R_{12}	5·6Ω, 3W, wire-wound.
R_{13}	33kΩ, 3W, wire-wound.
R_{14}	(1/64) × field resistance, Ω.
R_{15}	(1/32) × field resistance, Ω, wire-wound pre-set potentiometer.
R_{16}	470Ω, ¼W.
R_{17}	1kΩ, ¼W.
R_{18}	56kΩ, ¼W.
R_{19}	1kΩ, ¼W.
R_{20}	470Ω, 1W, high stability.
R_{21}	4·7kΩ, 12W, wire-wound.
R_{22}	2kΩ, wire-wound pre-set (variable).
R_{23}	5kΩ, speed control precision potentiometer.
R_{24}	1·2kΩ, ¼W, high stability.
R_{25}	330Ω, 1W (when D_{24} and R_{39} are included, reduce to 150Ω, ¼W).
R_{26}	2·7kΩ, ¼W, high stability.
R_{27}	3·3kΩ, ¼W, high stability.
R_{28}	82kΩ, ¼W, field compensation, adjust to suit requirements.
R_{29}	1kΩ, ¼W, high stability.
R_{30}	5·6kΩ, ¼W.
R_{31}	6·8kΩ, ¼W, high stability.
R_{32}	3·9kΩ, ¼W, high stability.
R_{33}	12kΩ, ¼W.
R_{34}	1kΩ, ¼W.
R_{35}	15Ω, ¼W.
R_{36}	120kΩ, ¼W, high stability.
R_{37}	6·8kΩ, ¼W, high stability.
R_{38}	6·8kΩ, ¼W, high stability.
R_{39}	68Ω, 1W, or zero Ω (see note).
R_{40}	560Ω, ¼W.
R_{42}	82Ω, ¼W.
R_{43}	220Ω, 1W.
R_{44}	100Ω, wire-wound pre-set.
R_{45}	1kΩ, ¼W.
R_{46}	10kΩ, ¼W.
R_{47}	50kΩ, carbon pre-set, potentiometer Mullard type E097AD.
R_{48}	220Ω, ¼W.
VDR_1	Asymmetrical v.d.r., Mullard E295ZZ/01.

(There are no components numbered C_6, R_8, R_9, R_{41})

Current Transformers
Output voltage 2V into 7·5Ω with full load primary current.

Note:
Without additional stabiliser voltage regulator diode D_{24}, and with $R_{39} = 0\Omega$, the set speed changes by about 2·5% for a 20% change in supply voltage. With the additional diode D_{24}, and resistor R_{39}, the set speed changes by only 0·25% for a 20% change in supply voltage. Both measurements are for maximum set speed.

APPENDIX 3

Components for Simple Control System (Fig. 15)

A. Mullard Thyristor Stack (D_1, D_2, D_9, CSR_1, CSR_2)	Current rating 2 × motor full load current.
C. Current-limiting module.	Mullard MY5051.
D. Trigger Module	Mullard MY5001.

Diodes

D_3, D_4	Mullard BYX22–80 (or to suit field current).
D_5, D_6, D_7	Mullard BYX10.
D_8	Mullard BZY88–C4V7.

Capacitors

C_1	4·0μF, 1000V, paper.
C_2	(10^6/550) × armature time-constant, in μF: typically 20μF, 50V, electrolytic.
C_3	250μF, 6·4V, Mullard 428AR/C250.

Resistors

R_1	5·6Ω, 3W, wire-wound.
R_2	33kΩ, 3W, wire-wound.
R_3	(1/full load armature current in amperes)Ω.
R_4	(1/7) × armature resistance, Ω.
R_5	22Ω, ¼W.
R_6	47Ω, ¼W.
R_7	500Ω, wire-wound (speed control).
R_8	3·3kΩ, 12W.
R_9	Thermistor, Mullard VA3201.
R_{10}	6·8kΩ, 12W (see MY5001 data).
R_{12}	200Ω, pre-set, ¼W.

REFERENCES

1. MERRETT, J., 'Thyristor Speed Control of D.C. Shunt Motors from a Single Phase Supply', **Mullard Technical Communications,** Vol. 8, No. 80, March 1966, pp. 298 to 315.

2. MERRETT, J., 'Instructions for Selecting a D.C. Motor for Thyristor Speed Control', **Mullard Technical Communications,** Vol. 8, No. 80, March 1966, pp. 316 to 327.

Articles on Power Topics

The following publications on power topics are available, free of charge:

	Reference
Power Control for Industry	TP840
Power Rectification with Silicon Diodes	TP552
Current Overload Protection in Thyristor Phase Control Systems	Ch
Regenerative Braking of D.C. Motors	Cg
Transient Performance of Heatsinks for Power Devices	Cd
Parallel Operation of Silicon Rectifier Diodes	Cc
Rectifier Diode Operation at Kilocycle Frequencies	Cb
Thermal Resistance of Semiconductor Devices under Steady-state Conditions	Bs
Transistorised Circuits for SCR Firing	Bp
Firing Requirements for Silicon Controlled Rectifiers*	Bf
Mains Overvoltages: Protection and Monitoring Circuits	Bc
Design of Cooling Fins for Silicon Power Rectifiers	Ak
Idealised Rectifier Circuit Performances	Mullard Technical Communications No. 82
Firing of Ignitrons by means of Thyristors	Mullard Technical Communications No. 82

and also Refs. 1 and 2 on page 103.

* * *

"Power Control for Industry" is concerned with rectifier stacks. The other items in the list are articles or reprints of articles which have appeared in *Mullard Technical Communications.*

Data sheets for Rectifier Diodes, Thyristors, and Stacks will be found in Volume 4 of the **Mullard Technical Handbook.** The Handbook is available on a subscription basis.

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Coil specifications: the specification for L_4 should be 300 turns, 80 turns/layer; for L_5, 400 turns, 100 turns/layer.

Line output transformer specification: the tappings for winding 1-9 should be 32, 48, 95, 145, 340.